zero at that station, or that they are as previously reported if no action to correct them has been taken in the meantime.

The matrix $E = e_{ij}$ may thus be equated to the matrix dC.A in (23), so we have dC.A = E or A'.dC' = E'.

This is an insufficient set of equations for the unknowns dC', but a unique solution exists for which the sum of the squares of the dC's is least, that solution being

$$dC' = A(A'A)^{-1}.E'$$

Moreover, if $A = C'(CC')^{-1}$ then $A(A'A)^{-1} = C'$, therefore dC' = C'.E' or

$$dC = E.C$$

Thus, given the original estimate of C and the residual correlation measurements reported by receiving stations, the error dC in the original estimate may be calculated and the estimate of C gradually refined.

As mentioned above, if the reverse SACCH signalling capacity does not allow all errors to be reported every time, it is sufficient to report only the largest. The transmitter can choose only to correct the largest there and then, or to wait until others are reported. In order to ensure that others are reported, the transmitter can request the receiver to make specific measurements via the forward SACCH channel. These refinements are mentioned for the sake of completeness in describing the scope of the invention, but the extra complexity is probably not needed in a satellite-mobile communications system where the relative positions of mobiles in the satellite beams changes only slowly relative to the speed of communications.

The control processor obtains initial estimates of the downlink C and A matrix coefficients, measured on the uplink by syncword correlation as described earlier, to the downlink frequency. The control processor then continually outputs corrected A-matrix coefficients suitably translated to the downlink frequency as described above to transmit matrix processor 1704.

A complication can arise in performing this translation due to phase mismatches between each antenna element channel. It was stated above that the relative amplitude between signals on the uplink and downlink

frequencies could reasonably be considered to be the same, and that the relative phase between signals can be scaled by the ratio of up- and downlink wavelengths. However, consider the case where phase mismatches exist between the channels that relay the mobile-satellite uplink signals from each antenna element. The signal phases are then not just antenna element phases, PHI(i), but contain the additive mismatch terms, dPHI(i). If PHI(i)+dPHI(i) is then scaled by the ratio of wavelengths, the PHI(i) part will scale correctly but the mismatch part dPHI(i) will not because there is no correlation between phase mismatches on the up- and downlink paths. If the up- and downlink phase mismatches are denoted respectively by uPHI(i) and dPHI(i) then we need to calculate:

a.(PHI(i)-uPHI(i)) + dPHI(i)) ; where a is the wavelength ratio.

This can be written a.PHI(i) + (dPHI(i) - a.uPHI(i)) and the term dPHI(i) - a.uPHI(i), which is at least a single constant, has to be determined in some way to translate the A- or C-matrix coefficients determined from receiving mobile signals to the coefficients that shall be used for transmitting to the mobiles. This can, for example, be done by a fixed system calibration that is carried out with the help of a few monitoring stations or "dummy mobiles" located at different positions throughout the service area. Alternatively, by having the mobiles also measure a limited number of residual correlations with signals other than their own, and report these correlations on the slow associated control channel (SACCH), the system can receive enough information to perform the necessary calibrations for phase mismatch continuously. Such reported information can also facilitate calibrating out amplitude mismatches if required.

The present invention can also be employed to improve the capacity of landbased cellular radiotelephone systems. Such systems generally employ 3-sector antennas to illuminate three adjacent cells from the same site, as described above. Because isolation between sectors is not high (in

fact isolation is almost zero for a mobile on the border of two sectors), it is not possible with conventional systems to permit use of the same frequency channel in all three sectors. According to exemplary embodiments of the present invention, however, the same channel can likely be employed as many times as there are antenna elements to form sectors. Thus a three-sector antenna (typically formed by three vertical collinear stacks of dipoles in a corner reflector) provides the opportunity to re-use the same channel three times.

Land-based cellular communication capacity is limited by the parameter of carrier to co-channel interference ratio (C/I). The C/I which would be obtained if signals on the same frequency are radiated around 360 degrees of azimuth is the same as the C/I which would be obtained with centrally illuminated cells. A 3-cell cluster or site then becomes the equivalent of a centrally illuminated cell as regards the re-use pattern needed to achieve a given C/I. It is known that a 21-cell re-use pattern is needed to provide the required C/I in the AMPS system, therefore a 21-site re-use pattern would be needed if all sectors in the same site used the same frequencies over. This compares with the 7-site, 3-sector pattern employed conventionally, showing that what has been gained from using the same frequency in every sector has been lost by the need to increase the re-use pattern size from 7 sites to 21 sites. Thus, according to this exemplary embodiment of the present invention three or more sectors or antenna elements around the 360 degrees of azimuth should be used.

Figure 18 shows an exemplary cylindrical array of slot antennas 1800 suitable for implementing the present invention in landbased cellular systems. The array consists of rings of eight slots around a metallic cylinder. Horizontal slot antennas give the desired vertical polarization, and the slots are a half wavelength long, e.g., approximately 16 centimeters for the 900 MHz band. It can be desirable to employ alternatively circular polarization at the base station combined with linear polarization at the mobile phone, especially when the mobile is a hand portable of uncertain

antenna orientation. Circular polarization can be formed by using crossed slots, crossed dipoles or a hybrid slot-dipole combination for the array elements. It is often convenient when using such structures to form both polarizations simultaneously, and this can be exploited by using opposite circular polarizations for transmitting and receiving to reduce transmit-receive coupling.

Element spacing around the cylinder must be somewhat greater than a half wavelength to avoid the slots from running into each other, although it is possible to stagger alternate slots by a small vertical displacement to reduce their potential mechanical interference or electrical coupling with each other. If, for example, 0.75 wavelength spacing is used, then cylinder circumference is 6 wavelengths, that is a cylinder radius of less than one wavelength or about one foot. Such an antenna is considerably smaller than conventional three sector antennas. A number of rings of such slots are stacked vertically with between, for example, 0.5 and one wavelength vertical spacing to provide the same vertical aperture and, therefore vertical directivity, as conventional cellular base station antennas. Slots that lie in a vertical column can be connected by feedlines 1801 that feed them in phase. The eight feedlines corresponding to the eight columns of slots are then connected to eight RF processing channels 1802. Each RF processing channel comprises a transmit-receive duplexing filter 1803, a linear transmit power amplifier 1804, an RF amplifier 1805, a downconvertor, IF filter, amplifier and A/D convertor 1806 for each frequency channel, and a corresponding transmit modulator 1807 for each frequency channel, the outputs of which are summed in summer 1808 before being amplified in power amplifiers 1804.

The digitized outputs for all eight columns of slots for each frequency channel are fed to a receive matrix processor 1809. The receive matrix processor 1809 is analogous to the matrix processor 650 of Figure 9. The matrix processor 1809 separates signals arriving on the same frequency but from different angles such that cochannel interference from mobiles in

communication with the same site is substantially suppressed. The separated signals are fed to voice or random access channel processors (not shown in Figure 18) analogous to channel processors 660 of Figure 9. Correlation measurements performed by the channel processors (not shown) are fed to a control processor (not shown) analogous to control processor 1702 in Figure 17. The control processor (not shown) produces both receive and transmit matrix coefficients for receive matrix processor 1809 and transmit matrix processor 1810 to produce a transmitted signal to every cochannel mobile in a non-interfering manner.

A difference in propagation conditions can arise in landmobile applications as compared to satellite applications, resulting in some modifications to the matrix processing that will now be described. Satellite propagation paths are substantially line of sight, and even if signal echoes from objects in the vicinity of the mobile occur, the sightlines from these objects to the satellite are substantially the same as the direct ray from the mobile to the satellite when compared with the relatively large cell diameters in satellite-cellular systems.

This is not true for landmobile systems. A substantial echo from a large building or mountain range on the other side of the antenna compared to the mobile can result in an echo that comes in from a direction anywhere between 0 and 180 degrees away from the direct ray. Since such echoes carry signal energy, it is often desirable to exploit them to provide a diversity path in the event that the direct ray fades or is shadowed in order to improve reception. Typically, the signal path from the mobile to the base station antenna consists of a number of rays caused by reflections from objects close to the mobile; these rays are received substantially from the same direction and combine to produce so-called Rayleigh fading. Since the base station antenna in, for example, large-cell applications is deliberately placed high at a good vantage point, there are not expected to be large reflecting objects in close proximity, for example within 1.5 Km, that could result in rays coming from substantially different directions. This means that

rays reflected from such objects and coming from an arbitrary direction would be expected to have traversed a larger distance, e.g., 3 Km, and thus suffered a delay of 10 μS or more.

To take care of both types of the aforementioned phenomena, that is a cluster of rays from substantially the same direction causing the signal to exhibit Rayleigh fading as well as a cluster of rays from a substantially uncorrelated direction representing a delayed signal, another term can be introduced into the receive matrix processing as follows.

A signal sample Si(t) received at the ith antenna element (column of slots) is the sum of non-relatively-delayed transmitted signals Tk(t) from mobiles k and signals relatively delayed by dt given by:

$$Si(t) = Ci1.T1(t) + Ci2.T2(t) \ldots\ldots + Cin.Tn(t)$$
$$+ Ci1'.T1(t\text{-}dt) + Ci2'.T2(t\text{-}dt) \ldots\ldots\ldots + Cin'.Tn(t\text{-}dt)$$

When the equations for all Si(t) are collected into matrix form, they can be written:

$$Sj = C.Tj + C'.T(j\text{-}m)$$

Wherein the suffix j of T means values at a current time and the suffix j-m means values m samples ago, corresponding to the delay dt. For example, if the signals are sampled every 5 μS, then for a delay dt = 10 μS, m would be equal to 2.

The signal fading of the undelayed ray can be considered to be due to varying C coefficients, with the transmitted signals T being constant, or the signals T can be considered to be varying due to Rayleigh fading and the matrix C to be constant. The latter is considered here, because after separating the fading signals T by using constant matrices, the voice channel processors can handle the fading signals as they do in landmobile systems.

If the signals T are considered to be fading, however, note that the fading on the delayed term is not correlated. In order to be able to consider T(j-m) as a delayed replica of the fading signals Tj therefore, the difference in fading must be explained by regarding the coefficients C′ as varying to convert the fading on the direct ray to the fading on the delayed ray. However, infinite values of C′ would then arise due to the varied coefficients being the ratios of Rayleigh fading values.

It is thus more convenient to regard the C-matrices as constant relative to directions of arrival, and to introduce an explicit set of Rayleigh fading variables to explain the fast fading. Each signal in the vector Tj, the first signal t1(j) for example, thus has an associated complex multiplying factor r1(j) representing the undelayed Rayleigh fading path from mobile 1 to the array. Assembling the factors r1,r2,r3...rn down the diagonal of a matrix, with zeros elsewhere and denoting this fading matrix by R0, the set of faded signals are then simply given by:

$$R0.Tj$$

Defining a different fading matrix R1 for the first delayed path, the delayed faded signals are given by:

$$R1.T(j\text{-}m)$$

Thus the signals out of the array elements are given by:

$$Sj = C.R0.Tj + C'.R1.T(j\text{-}m)$$

According to one aspect of the present invention, separation of the fading signals R0.Tj takes place using the separated signals R0.T(j-m) calculated m samples ago, based on the equation:

$$Ro.Tj = C^{-1}.\left[Sj - C'.\frac{R1}{Ro}.(Ro.T(j-m))\right]$$

It is seen that the previously separated signals R0.T(j-m) must first have their fading factors removed by division by R0 to replace the fading factors for the direct rays with the fading factors R1 for the delayed rays. This can cause numerical difficulties when a signal fades out completely so that its associated r-factor becomes zero. However, since the separated signal would also become zero, it is possible to assign a meaningful value to R0.T(j-m)/R0, using, for example, knowledge of the nature of the transmitted signal. For example, knowledge that the transmitted signal is a constant amplitude signal, or that it should be continuous between samples, could be used.

Alternate implementations and modifications of the principles set forth herein will be readily appreciated by those skilled in the art. For example, although the signal transmitted in any future cellular system will most likely be a digital signal, the principles of the present invention are also applicable to analog signals. In both cases, the fading spectra (i.e., the Fourier transform of a successive series of r-values) are narrowband compared to the modulation, which is the means by which information in the modulation can be distinguished from modulation caused by fading. In the case of digital signals, the modulators used at the transmitters are well characterized a-priori, so that the waveforms Tj that they will produce for a given information bit pattern can be predicted. If a known bit pattern is contained in a segment of transmission, a corresponding segment of the Tj waveforms can be predicted and correlating this with the received signals will yield an estimate of the corresponding T-value. This process is referred to as "channel estimation". The channel estimates may be updated after decoding each information bit. Due to the channel varying much slower than information bits and even more slowly than the sample rate of Tj, which may be, for example, eight times the information symbol rate, channel estimates are averaged over many successive samples of the T-waveforms, and are thus somewhat less noisy than the information signals themselves.

In the case of analog FM signals, for example, the modulation is known a-priori to be constant amplitude, varying only in phase. The rate of change of phase is known a-priori to be restricted to a value corresponding to the maximum frequency deviation, and the frequency variation is continuous and so the phase and at least its first and second derivatives are continuous. This a-priori knowledge can be used to predict a next Tj value from the previous history. For example, if Q_{ji} was the previous phase estimate and Q its derivative estimate, and A_j was the previous amplitude estimate, then $T_j = A_j$ EXP $({}_jQ_j)$ and $T_{j+1} = A_j$ EXP $({}_j(Q_j+Qdt))$. Hence, T_{j+1} is predicted from $T_{j+1} = T_j$ EXP $({}_jQdt)$.

Channel estimation techniques often use a Kalman filter including derivatives, in which a prediction of the next value of the channel estimate is made using an estimate of the time rate of change (derivative) of the signal, then the predicted channel estimate is used to predict the next signal sample point. The error between the predicted and received signal is then used to correct the estimate of the channel (the fading factor) and its derivative in such a way as to sequentially minimize the sum square error.

The same Kalman filter technique can also be used to estimate the diagonal elements of both R0 and R1. Having estimated these diagonal values, according to another aspect of the present invention, it is ascertained whether any value of R1 is greater than a corresponding value of R0. If a value of R1 is greater than a corresponding value of R0, that would indicate that the delayed ray is currently received at a greater strength than the direct ray. Then the column of C′ corresponding to that element of R1 is swapped with the corresponding column of C corresponding to R0 to form new matrices which are denoted by Cmax and Cmin. The greater element from R1 is swapped with the corresponding smaller element from R0 to form new R-matrices Rmax and Rmin, respectively. The elements of T(j-m) corresponding to the swapped R-elements are then swapped with the corresponding elements of Tj to form mixed vectors of delayed and undelayed signals denoted by Uj and Vj, respectively. The vector Uj can

contain some elements of Tj and some elements of T(j-m), while the vector Vj then contains the remainder. Thus, the equation for signals out of the array elements becomes:

$$Sj = Cmax.Rmax.Uj + Cmin.Rmin.Vj$$

This equation can then be solved to yield:

$$Uj = [Cmax.Rmax]^{-1}. \ [Sj - Cmin.Rmin.Vj]$$

Since each element of Rmax was chosen to be the greater of two, the chances of zero values are reduced. Furthermore, the Vj values that have to be subtracted from Sj are minimized by multiplication by Cmin, so if Vj values are wrong or noisy the error propagation into subsequent values will be attenuated.

The vector Vj, however, contains some as-yet uncalculated values. Assuming that the same elements of R0 and R1 are chosen for Rmax and Rmin next time, the as-yet uncalculated values of Vj belong to a future U-vector U(j+m). The previously calculated values of T contained in Vj come from a previous U-vector, U(j-m).

Cmin and Rmin can be partitioned into two matrices Cmin1, Rmin1 and Cmin2,Rmin2, the columns of which are associated with the Vj values that come from previous or U-vectors, respectively. Thus, the U-vectors can be described as:

$$Uj = [Cmax.Rmax]^{-1} \cdot \ [\ Sj \ - \ Cmin1.Rmin1.U(j-m) \ - Cmin2.Rmin2.U(j+m)]$$

The values of U(j-m) are known from a previous calculation, but the values of U(j+m) are not. Therefore, Uj is first calculated on the assumption that

all U(j+m) are zero. Then, m samples later when U(j+m) has been calculated on the assumption that U(j+2m) are zero, the calculated values of U(j+m) can be back-substituted into the above equation to give a refined set of values for Uj. These Uj values may be then back-substituted into a previous calculation of U(j-m) to refine that calculation, and/or forward substituted into the calculation of U(j+m), or both, to an iterative extent limited only by available processing power in the receive matrix processor.

Simplifying the above equation by denoting:

$$Ao = [Cmax]^{-1}$$

$$A1 = [Cmax]^{-1} \cdot [Cmin1.Rmin1]$$

$$A2 = [Cmax]^{-1} \cdot [Cmin2.Rmin2]$$

and substituting yields:

$$Rmax.Uj = Ao.Sj - A1.U(j\text{-}m) - A2.U(j+m)$$

If A1 has diagonal elements D1 and A2 has diagonal elements D2, then we can also write:

$$D1.U(j\text{-}m) + Rmax.Uj + D2.U(j+m) = Ao.Sj - (A1\text{-}D1).U(j\text{-}m) - (A2\text{-}D2).U(j+m)$$

The left hand side of the foregoing equation represents separated signals without cancellation of delayed or advanced rays. The separate channel processors can process these signals including delayed echoes to obtain better quality demodulation and decoding than if echoes had been subtracted. The improved decoded signals are useful in better producing the

required channel estimates. A device that can, for example, be used for this purpose is a Viterbi equalizer such as described in commonly assigned U.S. Patent Application No. 07/965,848, filed on October 22, 1992 and entitled "Bidirectional Demodulation Method and Apparatus", which is hereby incorporated by reference.

Thus, according to this exemplary embodiment of the invention, echoes of each signal are subtracted from estimates of other signals, but not from the estimate of the signal itself, to produce separation of signal+echo signals that are processed by individual channel processors. Echoes of each signal itself are left in additive combination with the signal and are used by a Viterbi equalizer. If echoes are not delayed or advanced by multiples of the modulation symbol period, a so-called fractional-spaced Viterbi equalizer can be used.

Such equalizers continuously estimate and update the amount and phase of additive echoes, as described in commonly assigned U.S. Patent No. 5,164,961 to Bjorn Gudmundson entitled "A Method and Apparatus for Adapting a Viterbi Algorithm to a Channel Having Varying Transmission Properties", U.S. Patent No. 5,204,878 to L. Larsson entitled "Method of Effecting Channel Estimation For a Fading Channel When Transmitting Symbol Sequences", and U.S. Patent Application Serial No. 07/942,270, filed on September 9, 1992 and entitled "A Method of Forming a Channel Estimate for a Time Varying Radio Channel", each of which are incorporated here by reference. The estimated values correspond to the diagonal elements of the diagonal matrices D1, Rmax, D2. Knowing Cmax and Cmin, Rmin1 and Rmin2 can then be determined, thus the channel adaptive equalizers in the individual channel processors can determine the Rayleigh fading functions R0 and R1.

A purpose for cancelling by subtraction cross-echoes, i.e., echoes of one signal that are additive to a different signal, is to provide separate signal sample streams that each depend only on one signal and its own echoes, as such can be handled by said channel-adaptive, Viterbi equalizers. For

completeness however, a further method will now be explained, that can be used when the number of signals to be separated is relatively few, for example, eight signals.

The receive matrix processor can be regarded as undoing the additive signal mixing that takes place in the aether. This is advantageous in simplifying the operation of the channel processors. However, as disclosed above, numerical difficulties can arise in dealing with signals that can periodically fade completely. This can result in certain matrices becoming singular, i.e., difficult to invert accurately. An equivalent problem arises in equalizers that attempt to undo the effect of a corruptive channel, for example a channel that suffers from selective fading that causes a null in the transmission function at some frequency. An inverse channel filter that attempts to undo the effect of such a channel would try to create infinite amplification at the null frequency, with consequent huge amplification of noise and other difficulties.

Therefore it is often proposed, as in the Viterbi equalizers cited, that the channel should not be "undone" by subjecting the received signal to an inverse channel filter to produce an undistorted signal that is then compared to the alphabet of expected symbols, but rather the alphabet of expected symbols is subjected to the same channel distortion as the signal by use of a mathematical model of the channel, and the distorted received signal compared with this predistorted alphabet.

According to a further exemplary embodiment of the present invention, a method is disclosed whereby no attempt is made to separate or "unmix" in the receive matrix processor the plurality of co-channel signals received by the array to produce separated signals that are then compared in separate channel processors with the alphabets of expected symbols. Instead, the alphabets of expected symbols are premixed in every possible way with the aid of a model of the mixing process that takes place in the aether, (i.e., with the aid of the C-matrix coefficients and the channel estimates R) and the

mixed alphabet is then compared to the mixed signals received by the array elements.

Such a scheme expands the number of possible mixed symbols in the alphabet exponentially according to a power of the number of signals. For example, suppose each signal is modulated with binary symbols. The expected symbol alphabet has only two symbols, 0 or 1. However, if the array elements receive weighted sums of eight signals, each of which instantaneously may be modulated with a 1 or a 0, the number of possible mixed signals that can be received is 2^8 or 256, if all the signals are aligned in time. If different signals are not time aligned, then a symbol period of one signal may overlap two symbols of another signal. Thus the waveform over a symbol period of one signal can depend on two symbols of each of the other signals. Nevertheless, each point of the waveform depends only on the one symbol of each signal whose symbol period it lies in. When echoes are taken into account, however, each waveform point can be dependent on two symbols of each signal thereby raising the number of possible values that can be observed to 65536. It will however be described below how, for example, a 256-state Viterbi algorithm can be used to jointly demodulate the signals from the array.

According to an exemplary embodiment aspect of the invention, and referring to Figure 19, a numerical machine has 256 sets of memory banks 1900 each associated with a specific 8-bit postulate for one previous binary bit in each of the eight signals, on which, due to a delayed echo, the received array signals will depend. The SMLSE controller 1910 now makes another 8-bit postulate 1920 for the current binary bit of each signal. How it makes this postulate is immaterial, as all postulates will eventually be tried. In the event that postulates are tried sequentially, they can, for example, be generated by an 8-bit counter. If however all postulates are tried in parallel using replicated hardware, each hardware unit would handle one fixed postulate that could then just be hard-wired in.

Together which each of the previous 8-bit postulates in turn plus the new 8-bit postulate, a set of eight signal predictors 1930 predicts the complex value of each signal incident on the array including one or more reflected rays by using the fading channel coefficients R and R′ and a-priori knowledge of the transmitted modulation or coding. The complex signal values are then combined in matrix processor 1940 by calculating the equation:

$$Sj = C.R.Tj + C'.R'.T(j\text{-}m)$$

where C and C′ are square matrices representing the directions from which the direct and delayed waves are principally received.

The calculated signals Sj are the signals that are expected to be received at the array elements if the hypothetical eight bits are correct. These hypothetical signals are then compared with the corresponding received signals R1,R2....R8 from the array elements using comparator 1950. Comparator 1950 evaluates the net mismatch of the eight predictions from the eight array signals by, for example, computing a sum of squares of the differences. Other means to produce a signal representative of the net mismatch are however known to the art, based on a mathematical expansion of the sum of squares, and can be used if considered advantageous for the particular implementation chosen. For example, note G. Ungerboeck, "Adaptive Maximum Likelihood Receiver For Carrier Modulated Data Transmission Systems", IEEE Trans. Commun. Vol. COM-22 No. 4, pp. 624-636, May 1974, U.S. Patent No. 5,031,193 to Atkinson et al., and U.S. Patent No. 5,191,548 to Bäckström et al., each of which is incorporated here by reference. The sum square error signal is fed back to the SMLSE controller 1910 which adds the error to the previous error stored in state memory 1900 against the previous 8-bit signal hypothesis 1921 employed in signal predictors 1930 to produce the signals ri'ti'.

The above procedure is carried out for each new 8-bit hypothesis in turn preceded by each of the stored, previous hypotheses. This results, for each new hypothesis, in 256 candidate cumulative error numbers depending on which preceding hypothesis was used. The lowest of these is selected to become the new cumulative error associated with the state corresponding to the new 8-bit hypothesis. When all possibilities for the new 8-bit hypothesis have been processed in this way, the state memory 1900 will contain 256 new cumulative error numbers associated with each new hypothesis, as well as a record of the best preceding hypothesis to each, i.e., that giving the lowest error, and the preceding hypotheses to those in turn, and so on. Thus each of the 256 states contains a candidate demodulated sequence of 8-bit values. The oldest values in these sequences will tend to agree and when this happens the machine is said to have converged to an unambiguous decision. The decided 8-bits are then extracted to yield one bit decision for each of the eight incident signals. If convergence does not occur and the sequence memory 1900 becomes full, the path history is truncated by believing the oldest byte of the state having the lowest cumulative error. That value is then extracted and the path history memories shortened by one byte.

The above process represents an alternative to attempting to separate signals that have been mixed by means of matrix processing. Instead, signals are hypothesized by models of the transmitters and models of the mixing process, and the hypothesis best corresponding to the observed, mixed signals is determined by the SMLSE machine 1910 in the manner described above. Thus, the need to invert a mixing process to separate mixed signals, which may be mathematically intractable, is avoided by instead applying the mathematically tractable mixing process to the hypothesized signals to predict the mixed signals that should be received by the array elements and picking the hypothesis that best matches the observed signals. This process will not fail when two mobiles using the same channel lie at the same bearing, the process then being equivalent to joint-

demodulation as, for example, disclosed in U.S. Patent Application Serial No. 08/__________ filed on November 22, 1993 and entitled "A Method and Apparatus for Joint Demodulation of CDMA signals with Multipath Time Dispersion".

The above-described exemplary embodiments of the present invention are applicable to satellite cellular communications systems to provide greater use of available bandwidth by permitting immediate spectrum re-use in adjacent cells. These techniques have also been described in relation to land cellular systems, where they permit, for example, re-use of the same frequency in adjacent sectors.

In practice, Inn both the satellite and land-based applications of the present invention, benefits are achieved by a combination of adaptive signal processing techniques linked to traffic management techniques. The traffic management techniques relate to continuously operational systems using TDMA or FDMA or a combination thereof in which calls are continually being terminated and new calls established. By selectively establishing new calls on time- or frequency-slots in such a way as to optimize a communications criterion, a natural sorting of traffic into groups using the same timeslot and/or frequency is established. The criterion relates to the ease with which the adaptive signal processing can separate signals on the same frequency and/or timeslot based on the reception of different, linearly independent combinations of them using a plurality of antenna elements.

According to yet another exemplary embodiment of the present invention, the signal processing does not adapt to the movement of mobile phones or to new call set-up and termination, but operates in a deterministic way and instead the traffic is adapted to the deterministic characteristics of the signal processing using a dynamic traffic channel assignment algorithm.

Conventional land-based cellular systems typically employ so-called sectorization, in which a single antenna mast carries three, 120-degree coverage antennas and illuminates three cells from a common site. This saves on real estate costs compared to illuminating the three cells using three

separate antenna sites at the cell centers. Six sector systems are also known. Cellular systems have conventionally employed analog FM voice transmission in which each conversation is assigned a separate pair of up- and down-link frequency channels respectively. TDMA systems are now being installed using digital speech transmission, in which each conversation is allocated a unique pair of timeslot-frequency channel combinations. In these conventional systems, however, the three, 120-degree sector antennas have the same radiation patterns for all frequencies and/or timeslots.

According to yet another exemplary embodiment of the present invention, rotationally offset radiation is provided between different frequencies and/or timeslots. For example, on frequency channel 1 the three, 120 degree sectors may be orientated towards 0 degrees (Due North), +120 degrees (South East) and +240 degrees (South West). On frequency channel 2, the three sectors may be orientated to 60 degrees (North East), 180 degrees (Due South) and 300 degrees (North West). In general one might have as many as 120 frequency channels with corresponding antenna sector patterns offset by only one degree from each other. Such a system cannot be implemented using today's fixed-beam cellular base station antennas, but can be arranged with using the exemplary cylindrically symmetric array and associated matrix processing of Figure 18.

Similarly, the antenna sector patterns can be rotationally staggered as between different timeslots in a TDMA system. In either the FDMA or TDMA or hybrid cases, this exemplary system determines at call set up, and optionally at regular intervals thereafter, the optimum time- and or frequency slot combination to use for communicating with the mobile station. The combination of a frequency and timeslot is abbreviated henceforth to simply "channel". The optimum channel is most likely one which has an associated antenna sector pattern pointing in the direction of that mobile. This channel would be selected if the selection criterion is, for example, maximum signal strength and the channel was free. If the criterion is maximum signal-to-interference ratio, different selections can result. Adaptive channel selection

methods can be used to implement the present invention as, for example, disclosed in U.S. Patent No. 5,230,082 to Ghisler et al. which is incorporated here by reference.

Figure 20 illustrates a set of staggered sector patterns that can be produced by the arrangement of Figure 18 using fixed matrix coefficients for each frequency channel of an FDMA system. Three lobes are created in this example on every frequency channel. The notation Pi(Fk) indicates the pattern of the i^{th} lobe on the k^{th} frequency channel. The matrix processing coefficients are preferably chosen such that P1(Fk) and P2(Fk) have minima where P3(Fk) has its maximum, and reciprocally. If the minima are zero, the three lobes are said to be orthogonal. That permits a mobile located in the nulls of Pl and P2 to receive maximum signal from P3 with no interference from the other two, which can thus carry separate signals. In general, true zeros will not be perfectly achieved, and the channel selection criterion will thus allocate a mobile to a frequency where the corresponding sector patterns result in maximum ratio of wanted signal to unwanted interference from other lobes and other cells. For example, the mobile M in Figure 20 would be allocated preferably F4, where the lobe P3(F4) has the maximum strength in the direction of the mobile M. If P3(F4) was not available, the next best allocation P3(F3) would be tried, and so on.

In practice, an FDMA cellular system such as AMPS has 1000 channels available, usually divided between two operators that handle a minimum of 400 each. Using the traditional frequency re-use pattern of 21, this results in around 20 frequencies being available in every cell or 60 per site. The angular difference between lobes on different frequencies would thus, in a three lobe system, be only 1/20th of 120 degrees or 6 degrees. In this example, different lobes at the same site all have different frequencies. Assuming uniform distribution of mobiles in angle, the channel allocation algorithm would result in each mobile being within a few degrees of beam center. This results in mobiles receiving better signals on average than in today's fixed sectorization patterns which, when optimized, are around 12dB

down at sector edges. If the wanted signal is improved in this way, the tolerance of interference from surrounding cells is improved such that the re-use pattern can be shrunk from 21 to a tighter re-use pattern such as 12, with a consequent capacity gain of 21/12. This can be achieved using the same number of lobes as sectors in today's cellular systems. If the number of lobes is increased to eight, as illustrated in Figure 18, a further 8/3 increase in capacity is obtained, to around five times current AMPS capacity. Moreover, allowing every cell to adoptively select any of the 400 frequency channels in attempting to maximize signal to interference ratio gains a factor of two in capacity relative to having a fixed subset of frequencies (1/21st or 1/12th of the total) in each cell. This is achieved when transmit power levels are also adapted to the varying radial distance of each mobile from its cell site. It is also possible to use all 60 site frequencies in each 120 degree sector by making lobes using the same frequency orthogonal, as defined above. Lobe separation is then 2 degrees and the channel allocation algorithm ensures not only that each mobile is within a couple of degrees of beam center, but also within a couple of degrees of the minima of the co-frequency lobes.

If instead of associating staggered sector radiation patterns with different frequency channels F1,F2,F3.... they are associated with different timeslots of a TDMA signal using a single frequency, the resulting radiation from the base station antenna will take a certain set of directions for timeslot 1, a set of rotated directions for timeslot 2 and so forth, such that the beams are apparently rotating with time. Thus in the TDMA context this exemplary embodiment of the present invention may be formulated in terms of creating beams which continuously rotate through 360 degrees over a TDMA frame, or more appropriately, rotate by 360/N degrees during a TDMA frame where N is the number of sectors of frequency re-use, and the data modulation for the next frame is shifted back one sector between successive frames such that data for the same mobile continues to be radiated in the same direction. Data destined for a particular mobile is indicated in

the US IS54 TDMA system by inclusion of a "Digital Voice Color Code" (DVCC) in TDMA bursts. Thus, for example, this technique can be described more simply in terms of rotating the antenna sector patterns in one direction while rotating the DVCC in the reverse direction at the same rate such that the same DVCC continues to be radiated in the same direction on successive frames.

Both exemplary FDMA and TDMA embodiments of the present invention provide mobile stations with the capability to determine coarse geographic position. In the FDMA version, the mobile measures relative signal strength on different frequencies. The frequency on which the greatest signal strength is received indicates the bearing of the mobile within a sector. The sector is determined by decoding sector ID information contained within the transmission.

In exemplary TDMA embodiments, the mobile does not even have to change frequency. The mobile instead notes the cyclic signal strength variation during a TDMA frame and then determines the peak and trough signal strength positions relative to timeslot 1, which can be identified by the slot ID information carried in each slot.

The cyclic signal strength variation can be processed over several cycles with the aid of a Fourier transform and the phase of the fundamental component relative to timeslot 1 will then indicate the mobile's bearing. Bearings from two base stations of known positions then fix the mobile position. The mobile can report the timing of signal strength peaks and the network can perform the position calculation, rather than the network having to send coordinates of base stations to the mobiles. Upon allocating a traffic channel to a called or calling mobile, the network can then determine the best of all available timeslot/frequency combinations to use.

The above-described can also be adapted to provide advantageous communications between mobile stations and an orbiting satellite. According to this embodiment, the antenna array signal processing is not adapted to various mobile positions, but rather mobiles are allocated to a specific

antenna array signal processing channel based on position in such a way as to optimize communications. That is, mobiles are adaptively allocated to communicate using one of a number of fixed, staggered antenna beam positions instead of adaptively steering the antenna beams onto the mobile positions.

The operation for satellite use may be modified slightly. The notion of fixed antenna beams would be applicable to a geostationary satellite, but may not be applicable to, for example, a low-orbit satellite that changes position relative to the earth. Then the position of a beam relative to a given mobile would move due to satellite motion if not due to mobile movement. If the satellite beams move over the earth relatively slowly in comparison with the average 3-minute call duration, it may be sufficient to allocate a mobile to a beam at call set up, as one would in the Geostationary case. However, according to this examplary embodiment of the present invention, beam directions can be adapted to remove the systematic motion of the satellite over the earth so that the area illuminated by each beam is static from satellite rise to satellite set. In this way, a mobile may remain allocated to the same beam irrespective of satellite movement during this period.

Furthermore, such a system of low-orbiting satellites would generally be arranged to provide continuous coverage whereby as one satellite sets, another rises. For example, it can be arranged that a satellite rising in the west takes over the illumination of the same area just being vacated by a satellite setting in the east. Then, as the adjacent area to the east of this experiences loss of the setting satellite, the rising satellite creates a new beam to dovetail in while having maintained the first beam over the original area, and so on until the new satellite has taken over illumination of all areas originally served by the setting satellite.

Thus, the application of this exemplary embodiment of the present invention in the case of moving satellites allows the illumination patterns from the satellite antenna to be compensated for satellite motion so as to

illuminate fixed areas of the earth while adaptively allocating mobiles to illumination patterns using a channel assignment algorithm that optimizes a communications quality criterion. This contrasts with mechanical methods of compensating for satellite motion by tilting the satellite or antenna so as to maintain the center point of at least one area constant. This mechanical method, however, cannot maintain the center points of cell illumination areas constant due to these areas changing shape from circular as the satellite moves overhead to elliptic and finally to parabolic at earth-edge illumination. Alternatively, this exemplary embodiment of the present invention can employ both a mechanical method for coarse compensation plus the method of adaptive antenna array signal processing to correct the illumination patterns for shape change as the satellite moves. Alternatively, signal processing can be used to hold the areas served by a particular frequency and/or timeslot constant while progressively creating new areas forward of the satellite's ground track that are being vacated by a setting satellite and while terminating illumination of areas to the rear of its ground track that are being taken over by a rising satellite.

The operation of this exemplary embodiment of the present invention is depicted in Figures 21(a) and 21(b). At a certain time T (Figure 21(a)) a rising satellite 2100 illuminates areas with frequencies (left to right) F1,F2,F3,F1,F2,F3,F1,F2 and a falling satellite 2102 illuminates further areas with frequencies (left to right) F3,F1,F2,F3,F1,F2,F3,F1 which continue the frequency re-use sequence. At, for example, time T+5 minutes (Figure 21(b)) the rising satellite 2100 has ceased to illuminate the rearmost Fl area 2104 which is presumably now obscured from view (i.e., the satellite is too low on the horizon for good communications with this area) while the setting satellite 2102 has stopped illuminating its rearmost area 2106 with frequency F3 for the same reason.

On the other hand, the rising satellite has created a new illumination area 2107 forward of its ground track to fill in the area vacated by the setting satellite. The rising satellite 2100 can appropriately illuminate the

new illumination area 2107 with the same frequency as its predecessor used. Meanwhile, the satellite 2102 that is setting with respect to this area is rising as viewed from areas ahead of its ground track, and uses the released capacity to create a new area 2108 forward of its ground track illuminated with frequency F2.

It will be appreciated that instead of different frequencies, the overlapping areas could have been allocated different timeslots in a TDMA frame, or different frequency/timeslot combinations in a hybrid FDMA/TDMA system. Either way, the availability of a large number of channels allows the overlapping beams to be much more finely spaced than in the example of Figures 21(a) and 21(b), so that it is almost equally effective to allocate a mobile to an adjacent beam as to the optimum beam. Logically, one should allocate a mobile preferably to a beam in which the mobile is centrally located. However if the corresponding channel is occupied, the mobile can be allocated to a slightly off-center beam and may be handed over to the on-center beam when the call using that channel terminates.

In the exemplary TDMA embodiment, the rising satellite and the setting satellite can both illuminate the same area using the same frequency, providing different timeslots are used. Thus, a channel and satellite allocation strategy according to the present invention is to allow calls in the changeover region to terminate naturally on the setting satellite and to re-employ their vacated timeslots in the same region and on the same frequency to set up new calls using the rising satellite.

Figure 22 is a block diagram of an exemplary control processor that supplies the matrix coefficients to the numerical matrix processor of the hub station, e.g., block 1603 in Figure 13. Inputs to the control processor 2200 include satellite orbital data including attitude control information from the independent satellite Telemetry, Tracking and Command (TT&C) subsystem (not shown). Using satellite orbital and satellite antenna pointing information (attitude control information) and an input from a real time

clock, the control processor 2200 can determine the matrix coefficients needed such that a given area will be illuminated by a specific frequency in a specific time slot. These coefficients are systematically updated in step with changes in the real time clock to maintain these illuminated areas approximately fixed irrespective of satellite motion. The control processor 2200 also receives information transmitted from mobile stations making a random access on a calling channel that allows the control processor to determine the best available channel/beam combination to use. This information provides a rough indication of the location of the mobile and the control processor then determines the available beam centered most closely on this location. This in turn determines the frequency and/or timeslot that should be used for communication with the mobile.

It will be apparent to one skilled in the art that TDMA and FDMA are not the only access methods that are compatible with the present invention. Code Division Multiple Access (CDMA) can also be used, where illumination areas are similarly staggered over the earth according to a CDMA code use pattern. Indeed, any multiple access method which defines a channel by means of a set of access parameters can have systematically staggered illumination areas depending on those access parameters. Moreover, the access method used on the downlink can be different from the access method used on the uplink, providing a set of uplink access parameters is paired with a corresponding set of downlink parameters in each offset beam or staggered illuminated area. For example, a combination of TDMA on the downlink with CDMA on the uplink, the uplink transmission being continuous apart from a short interruption during reception of the downlink slot, is disclosed in U.S. Patent Application Serial No. ________________ filed on January 11, 1994 and entitled "TDMA/FDMA/CDMA Hybrid Radio Access Methods", which is incorporated here by reference. Having described an exemplary embodiment wherein dynamic traffic channel assignment allows traffic to be adapted to the deterministic characteristics of signal processing, a complementary,

exemplary embodiment will now be described wherein capacity can be optimized through coding and frequency re-use schemes.

The ultimate capacity of a cellular satellite communications is limited by available bandwidth, as power limitations can always be solved by money, e.g., by launching more satellites. Practically, however, there are financial constraints on power and political constraints on bandwidth, therefore it is desirable to use bandwidth efficiently without significant sacrifice of power efficiency.

It shall be appreciated that the trade-off of bandwidth and power efficiency for a cellular (i.e., area or global coverage system) is different from that of a single link, as a single link trade-off does not consider the possibility of frequency re-use in adjacent cells. The units of capacity in the two cases are in fact different, being Erlangs/MHz for a single link and Erlangs/MHz/SqKm for a cellular system.

A cellular system illuminates a service area by dividing it into cells and using some fraction 1/N of the total available bandwidth in each. A cluster of N neighboring cells can thus be allocated different 1/N fractions so that they do not interfere. Outside the cluster, for cells far enough away, the bandwidth can be re-allocated to another cluster.

The reduction of interference by employing an N-cell re-use pattern is measured in terms of the carrier to interference ratio C/I, which is the ratio of wanted signal power to the sum of the power of all unwanted spectrally and temporally overlapping signals. Increasing N increases the C/I, but reduces the bandwidth available in every cell, thereby limiting system capacity. Reducing N worsens C/I but increases the bandwidth available to every cell. If the modulation and coding scheme can tolerate the reduced C/I, capacity will thus be increased by reducing N.

One method of providing greater C/I tolerance is to use redundant coding. This method increases the bandwidth per signal, however, which offsets the benefit conferred by shrinking the re-use pattern N. The question to be asked is where the optimum lies.

In land-based cellular systems, this question has been deeply studied, leading some people to conclude that the extreme bandwidth expansion of CDMA techniques combined with immediate frequency re-use in every cell provides the highest capacity. According to exemplary embodiments of the present invention, however, it is found that capacity increases with increasing coding and reduction of N until N=1 is reached with a coding rate of about 1/3 (for landcellular). At this point the system is not regarded as being truly CDMA as each channel is still only used once in every cell. CDMA can be defined as the use of each channel more than once in each cell, i.e., a fractional value of N. For example, N=1/2 means each channel is used twice in every cell, which would be classified as CDMA.

Whether this further reduction of N to fractional values continues to increase capacity depends on what type of CDMA system is employed and on the nature of the propagation channel and receiver complexity used in the system.

Three types of CDMA systems may be distinguished:

i) Conventional, non-orthogonal CDMA

ii) Orthogonal CDMA

iii) Interference cancellation CDMA (subtractive CDMA, joint demodulation, etc.)

For the landbased cellular world, it is found that the capacity drops off below N=1 for CDMA of type (i), levels off for orthogonal CDMA (which is really equivalent to giving every signal a unique frequency or timeslot) and increases for systems of type (iii). Moreover, the gain found in systems of type (iii) for landbased cellular where N<1 is due to the high near-far environment such that the interference averaging inherent in CDMA techniques includes many transmitters of significantly reduced power, and due to the landbased cellular scenario being C/I limited rather than noise, C/No, limited. Neither of these features are relevant to satellite communications systems. Accordingly, the present invention explores what

kind of coding/frequency re-use trade-off would maximize capacity for a given bandwidth allocation in C/No-limited satellite communication systems.

The signal spillover between cells in a landbased cellular system is a function of the fourth-power-of-distance propagation law. In cellular-satellite systems C/I is a function of antenna beam pattern sidelobes. It is necessary therefore to develop some model of antenna beam patterns to perform coding optimization.

The beam pattern of the antenna depends on the surface current distribution over the aperture, called the aperture illumination function. Without invoking the supergain phenomenon, the most efficient use of aperture is obtained with uniform illumination. This gives the best gain but the highest sidelobes. The radiation pattern is plotted in Figure 23 for a uniformly illuminated circular aperture. The sidelobes in the E and H planes are slightly different owing to an extra cosine factor that appears in the plane containing the surface current vector. This difference manifests itself as cross-polarization components when circular polarization is employed. Henceforth the E and H plane patterns will simply be averaged for the calculation of C/I.

Reference is made again to Figure 5 which illustrates a 3-cell frequency re-use pattern, wherein the shaded cells use the same channel f1 while the others use f2 or f3. This re-use pattern will be used to investigate the coding/frequency re-use tradeoff for satellite communications, however, those skilled in the art will appreciate that any re-use pattern, e.g., 7, 9, 12, 21, etc., could be used. Interfering cells lie on the points of a hexagon and it suffices to consider the first two rings of six interferers. Before their interference levels can be computed however, it is necessary to choose the correct scaling of the beam pattern to match the cell diameter.

If the beams are scaled to cross at -3dB relative to peak gain midway between two cells, it is well known that this does not result in maximum beam-edge gain. A higher gain is achieved if the beam is narrowed, which increases the peak gain more than increased edge-loss experienced.

Figure 24 is a plot of peak gain 2401 (at the center of a cell), edge gain 2402 (midway between two cells) and the gain midway between three cells 2403 as a function of the two-cell crossover point in dB down from peak, relative to the peak gain of the -3dB crossover case. For reasons that will be explained later, the two-cell edge gain has been scaled by a factor of two in this plot (i.e., 3.01dB is added) and the 3-cell edge gain has been scaled by a factor of 3 (i.e., 4.771dB has been added). This does not affect where the respective gains peak, but affects the perception of which of the three is the worst case. According to this graph, the worst case occurs midway between two cells, and the worst case gain is maximized when the 2-cell edge is 3.8dB down on the peak gain, i.e., at point 2404.

The way in which the C/I parameter depends on the beam crossover point is shown for the 3-cell re-use pattern of Figure 5 in Figure 25. Figure 25 is plotted as a function of mobile station distance from beam center for crossover points of -3, -3.5 and -4dB showing that C/I over most of the cell radius increases if the beams are narrowed beyond that which gives maximum edge gain. If necessary, choosing a crossover point of -4.5dB would cause negligible loss of edge gain while improving C/I at cell center by a further 3dB to about 20dB. C/I at cell edge according to Figure 4 would be just less than 10dB, but this includes the unmitigated beam edge crossover loss which, as will be explained later, will not be incurred because no mobiles need be located there.

If mobiles assigned to a particular channel and beam are chosen to be those located within 25% of the maximum cell radius, the C/I for all points within that area will be as plotted in Figure 26. The worst case C/I is maximized to about 23dB with a beam crossover design point of -5.5dB, somewhat beyond that which gives maximum edge gain, so in practice the -4.5dB crossover point would be used, giving a worst case C/I of 18dB.

The same calculations are now repeated for the N=1 frequency re-use pattern, i.e., immediate frequency re-use in adjacent cells, and results are plotted in Figure 27. This shows a cell-center C/I of 14dB for the -4dB

crossover case, but a cell edge C/I of about -1.5dB. The thickness of the curves in Figure 27 is due to superimposition of plots for all mobile angular positions in the cell, and dependence on angular position is a little more noticeable in the N=1 case than the N=3 case. The first 6 rings of 6 interferers were summed to obtain the plot of Figure 27.

Again, as will be shown later, mobiles using a particular channel and beam can be restricted to those lying within 25% of beam center or less, so it is of interest to maximize the worst case C/I within this region, as shown in Figure 28. The worst case C/I in this case maximizes at 13dB by choosing the beam crossover point to be -4.8dB, but this may be restricted to -4.5dB to avoid loss of beam edge gain for only a small reduction in C/I to 12.5dB.

Figures 29-34 give the results for repeating the whole process described above for a different aperture illumination function, the 1/2-cosine wave. This aperture illumination function is slightly less aperture-efficient than a uniform distribution, but gives lower sidelobe levels (see Figure 29) leading to higher C/I, particularly in the 3-cell re-use case (20dB over whole cell, or 27dB out to 25% of radius). As seen in Figure 34, the C/I for immediate frequency re-use out of 25% of cell radius is 13.5dB with the practical beam-edge crossover point of -4.5dB. Since this was 12.5dB for the uniform illumination, case of Figure 28 it should be noted that this value is not very sensitive to the aperture function being used.

The bit error rate is generally plotted as a function of Eb/No, which is equal to the ratio of signal power to the noise power if it were to be measured in a bandwidth equal to the bit rate. The latter does not imply any assumption that any physical receiver filter bandwidth must be equal to the bitrate; it is only that "bitrate" is a convenient unit of bandwidth for defining the noise density with which the performance of any given receiver will be tested. The receiver error rate performance will of course depend on the choice of its bandwidth, and that which optimizes performance at a given

Eb/No may be greater or less than the bitrate depending on the modulation and coding being used.

The C/I parameter is, by contrast, the ratio of wanted to unwanted signal power in the physical receiver bandwidth. This ratio, however, is independent of the choice of receive filter if the C and I have the same spectral shape and are thus equally affected by the filter. With the simplification that any 'I' passed through the receive filter will have the same effect on error rate as an equivalent amount of white noise, NoB, passed by the filter, where B is the noise bandwidth, the effect of I can be expressed in terms of an equivalent increase in noise density No by an amount Io to No+Io, where Io is given by

$$I = Io.B \qquad \text{i.e.} \qquad Io = I/B$$

For BPSK modulation, the optimum receiver bandwidth is indeed equal to the bitrate, while for QPSK modulation the optimum receiver bandwidth is equal to half the bitrate. The bitrate here though is the coded bitrate/chiprate, whereas the bitrate for defining Eb/No is the information rate. Thus:

B = Bitrate/r for coding rate r in the BPSK case,

B = Bitrate/2r in the QPSK case, and for general M-ary modulation,

B = Bitrate/$rLog_2(M)$ = Bitrate/mr where m is the bits per symbol.

Therefore the total bit energy Eb to noise plus interference density ratio is given by:

$$\frac{Eb}{(No+Io)} = \left[\frac{No}{Eb} + \frac{Bitrate}{B}\cdot\frac{I}{C}\right]^{-1} = \left[\frac{No}{Eb} + mr\cdot\frac{I}{C}\right]$$

For less than 0.5dB degradation of the Eb/No required for a given error rate due to finite C/I, the value of mr.I/C should thus be one tenth No/Eb.

For example, if the ratio Eb/No without interference of 3dB is desired, then to operate at 3.5dB Eb/No the C/I must be 10mr.Eb/No. For BPSK or QPSK and different code rates, the required C/I for exemplary coding rates is given below:

REQUIRED C/I using	BPSK	QPSK
Coding rate 1 (none)	13.5dB	16.5dB
1/2	10.5dB	13.5dB
1/3	8.7dB	11.7dB
1/4	7.5dB	10.5dB

The above is for a static channel and does not take into account that lower Eb/Nos are needed with lower rate codes for the same error rate.

"Error Correction Coding for Digital Communications" by Clark and Cain gives the required Eb/Nos for 0.1% BER for constraint length 6 convolutional code rates of 1, 3/4, 2/3, 1/2 and 1/3 as follows:

r	Eb/No for BER = 0.1%
1	6.7dB
3/4	3.9dB
2/3	3.5dB
1/2	3.0dB
1/3	2.6dB

By extrapolation it may be estimated that rate 1/4 would require 2.3dB with diminishing returns thereafter. Using these Eb/No figures, the C/I required for less than a given degradation are calculated to be:

	REQUIRED C/I for 0.5dB		1.0dB loss		3.0dB loss	
	BPSK	QPSK	BPSK	QPSK	BPSK	QPSK
Coding rate 1 (none)	17.2dB	20.2dB	13.7	16.7	9.7	12.7
3/4	13.2	16.2	10.9	13.9	6.9	9.9
2/3	12.2	15.2	8.7	11.7	4.7	7.7
1/2	10.5	13.5	7.0	10.0	3.0	6.0
1/3	8.3	11.3	4.8	7.8	0.8	3.8
1/4	6.8	9.8	3.3	6.3	-0.7	12.3
1/5	5.7	8.7	2.2	5.2	-1.8	1.2dB

Thus, while the Eb/No for a given error rate levels out with increasing coding, the C/I required continues to decrease due to the continually increasing bandwidth. This equates to the separate concepts of coding gain (which applies to Eb/No) and processing gain (which applies to C/I). Coding gain is bounded by Shannon's limit, while processing gain continues to increase with bandwidth as in a CDMA system.

The above results for the static channel are pessimistic for fading channels. When Rician or Rayleigh fading is present, the mean Eb/No can be increased above the static Eb/No requirement to maintain the same error rate. However, on the satellite downlink the C/I does not exhibit fading, because both the I and C reach a given mobile over exactly the same channel and fade by exactly equal amounts. Thus the C/I does not decrease by 10dB when the Eb/No fades 10dB, but instead stays at the original value.

In the fading channel, since the error rate at the mean Eb/No is considerably less than the target value, and when it fades to the static Eb/No value still only equals the target value, it is clear that the error rate only reaches the target value by virtue of fades to below the static Eb/No value. In fact it can be shown that the preponderance of errors arise from the instantaneous Eb/No region well below the static Eb/No value, where the same C/I causes less degradation. It can be assumed that lower C/I values

can be tolerated in conjunction with the higher Eb/No values needed to account for fading.

Thus the 12.5-13.5dB C/I values achievable out to 25% of beam radius with immediate frequency re-use are acceptable with coding rates of 1/2 to 1/3 and using QPSK. Increasing the re-use pattern to N=3 would yield 3 times less bandwidth per cell, requiring that all coding be removed and even higher order modulations than QPSK to be contemplated in order to achieve the same bandwidth efficiency, but with the penalty of needing considerably higher power (e.g., Eb/No of 7.7dB to 10.7dB for achievable C/I's with the 3-cell re-use pattern). Thus there is no gain in bandwidth efficiency using an N=3 or greater frequency re-use pattern instead of N=1, only a major penalty that either is paid in power efficiency (for maintaining the same bandwidth efficiency by removing coding, as in the AMSC system) or in bandwidth efficiency if coding is retained.

To make use of the above result it is explained below how use of a beam can be restricted to mobiles located, for example, only out to 25% of the beam radius or less.

A gain of up to 2:1 in capacity can be achieved in cellular systems with a type of frequency planning known as "re-use partitioning". In a simple form of re-use partitioning, the available channels in a cell are partitioned into three sets that are preferentially used for a) mobiles within the inner 1/3 of the total cell area; b) for mobiles between 1/3 and 2/3 of the cell area, and c) for mobiles in the outer 1/3 of the cell area. Assuming a uniform area distribution of mobiles within the cell, this partitioning achieves equal demand for each of the channel sets. The allocation of channels to equal area rings is then permuted in the neighboring cells according to a 3-cell re-use pattern with the result that no two neighboring cells use the same channel out to their mutual border, with consequent increase in C/I for no loss of capacity. The overall re-use pattern to achieve a given C/I may then be shrunk to achieve an increase in capacity. Based on the foregoing principles, re-use partitioning and coding can be optimally combined

according to examplary embodiments of the present invention which will now be described.

Figure 35 shows a simplified example form for the case where three channels or groups of channels (which may be frequencies, timeslots, codes or combinations thereof), designated by the colors black, red and green are available in every beam. The beam edges at the design crossover point (e.g., -4.5dB) are shown by the larger colored circles of Figure 35.

The large black touching circles thus refer to beams using the "black" channel and touch at -4.5dB down from the peak gain. The large red touching circles represent the beam patterns for the red channel. These are displaced relative to the "black" beams, and this fixed displacement is achieved for example by modifying the phases of a phased array for the "red" channel relative to the "black" channel. It may also be achieved by use of a multiply-fed parabolic reflector, in which the unmodified beam patterns are used for the "black" channels, but in which three adjacent feeds are energized each with 1/3rd of the energy destined for a "red" cell. Due to coherent addition, the gain in the center of the "red" cell will be 3 times a "black" beam gain at that point, effectively "filling in the hole". The "green" beams are formed in exactly the same way for the "green" frequency or timeslot. This is achieved using ground based hardware that directs the appropriate combination of signals through the transponder channel directly associated with each of the feeds.

In Figure 35, the smaller circles show the areas out to which a particular channel is to be used, beyond which a different channel is available with a more centrally directed beam. The area has been filled in in the case of the black beam to assist in identifying it. This area extends from the center of a beam out to the beam radius over root(3), that is the "cell" area is only 1/3rd of the "beam" area, and mobiles in the "cell" only use the "beam" out to slightly more than 50% of the beam radius.

In practice, of course, many more than three channels are available per beam, so it is possible to plan for cells that are only 1/M of the beam

spot area where M is the number of channels available. If M=7 for example, beams are only used out to 1/root(7) of their radius, as illustrated in Figure 36. In practice M will be at least 100, so cell radius can be 1/10th of the beam radius, hence the gain and C/I performance of the beam configuration is only important for a fraction of the beam spot coverage. This does not necessarily mean it is possible to shrink the spot to obtain more gain, as it would not then be possible so easily to "fill in the holes". With a large number of offset beams it is desirable to phase the physical feeds to produce peak gain anywhere, and as indicated by Figures 24 and 30. The hardest place to obtain gain (by phasing only two feeds together) is midway between two spots, and the gain under those circumstances is maximized by choosing the beam edge crossover points as indicated on Figures 24 and 30. The gain between two beams is then twice the beam edge gain (e.g., 3dB up) while the gain between three beams is three times the gain of one beam at that point. This explains the scaling used on Figures 24 and 30 for comparing the gain at those three points.

Thus, the error correction coding of rate between 1/2 and 1/3 which is, in any case, contemplated for power efficiency reasons, can also provide tolerance of the C/Is obtained with immediate frequency re-use in every beam, if the technique of re-use partitioning just explained above is employed. The technique of re-use partitioning achieves this without resort to null-creation or interference cancellation, i.e., all antenna degrees of freedom are used to maximize gain. The technique of interference cancellation or creating pattern nulls at the center of neighboring cells can be employed as a further bonus to reduce C/I from neighboring beams to negligible proportions.

An exemplary coding scheme which can be used to implement this exemplary embodiment of the present invention is punctured convolutional coding based on rate 1/4 or 1/5th, but in which the coding of each uncoded speech bit is adapted according to its perceptual significance to a level between, for example, rate 1/2 and rate 1/5. Although BPSK tolerates 3dB

lower C/I than QPSK, there seems no reason to incur the 2:1 bandwidth efficiency loss. The C/I tolerance of QPSK with twice the coding is in fact better than BPSK with half as much coding, therefore a quaternary modulation can be used at least for the downlink.

The above discussion is based on coherent demodulation performance, which is achievable in the satellite-to-mobile channel using quite wideband TDM and is not achievable with narrowband FDM. The criterion for the downlink method is that the number of information bits to be demodulated and decoded shall be large over the time during which the fading component of the channel can be considered static, that is about 200uS at 2.5GHz and a vehicle speed of 100Km/Hr. Thus the information rate needs to be a couple of orders higher than 5Kb/s and with, for example, a 1/3 mean coding rate the transmitted bit rate needs to be greater than 1.5Mb/s, which will pass through a 1 MHz bandwidth channel using quaternary modulation. The capacity provided by systems based on the foregoing techniques is of the order of 100 Erlangs per MHz per spot area using a 4Kb/s vocoder or 166 Erlangs per MHz per spot using a 2.4Kb/s vocoder.

One exemplary technique for implementing the above-described re-use scheme of the present invention is by means of ground-based beam-forming, as described earlier in this specification. This involves providing feeder links to carry the signal for each antenna feed from the central ground station to the satellite in such a way that the relative phase and amplitude differences between the signals is preserved. Using such a coherent transponder, only one transponder channel is needed on the satellite per antenna feed point.

An alternative means of implementing the present invention for the fixed-beam case disclosed is set forth below, which avoids the need for coherent feeder links at the expense of more hardware on the satellite. The use of on-board beam forming is simplest if it is fixed and not variable, which may be more suited to a geostationary satellite that illuminates fixed

areas. For a non-geostationary satellite this exemplary embodiment of the present invention can still be employed, but it is not then so easy to obtain the advantage of systematically adapting the beam forming to compensate for satellite motion so that beams illuminate fixed areas.

The implementation of the fixed beam-forming transponder is shown in Figure 37 for the FDMA case, i.e., the total available bandwidth is divided into N sub-bands, each of which is used to illuminate areas on the ground according to a cellular re-use pattern such as shown in Figure 35. The case of three sub-bands - designated black, red and green as per Figure 35 - is used by way of illustration.

A set of transponder channels 37 receive signals from a corresponding set of feeder links and downconvert them to a suitable intermediate frequency for amplification and filtering. The I.F. outputs of 3710 are applied to I.F. beam-forming network 3720 which forms weighted combinations of the I.F. signals. The "black" channels are arbitrarily chosen to correspond directly to unmodified antenna patterns, i.e., black signal 1 shall be radiated directly and only through antenna feed number 1; black signal 2 shall be radiated only through antenna feed number 2, etc. The beam-forming network thus connects black channels with unity weighting into only that summing network corresponding to the designated antenna feed.

The red channels and green channels however shall be radiated with a beam pattern centered midway between three black beams. The red beam that shall lie midway between black beams 1, 2 and 3 is thus connected to their associated three summing networks through voltage/current weightings of 1/root(3) (power weightings of 1/3). One third of the "red" energy is thus radiated through each of the three feeds surrounding the desired "red" center. Likewise, the green beam lying midway between black beams 2, 3 and 4 is connected via weightings of 1/root(3) to the summers associated with feeds 2, 3 and 4. The weightings quoted above are exemplary and simplified for the purposes of illustration. Since the I.F. beamforming

network can in principle be realized with a network comprising mainly simple resistive elements, more complex sets of weights can be used with acceptable complexity impact. For example, a beam can be formed by feeding more than three adjacent feeds, and negative weights can be used to create nulls in the radiation pattern at desired places or otherwise to reduce the sidelobe levels in order to increase the C/I.

One method to form a resistive I.F. beamforming network uses a continuous sheet or thin film of resistive material deposited on an insulating substrate. This sheet is notionally regarded as corresponding to the two dimensional surface to be illuminated by the beams. Signal currents corresponding to the "black" beam signals are injected into the sheet at points disposed in correspondence to the centers of the "black" cells, while "red" and "green" signal currents are injected at sets of points midway between the black signal injection points and each other, as per Figure 35. Figure 38 illustrates the injection points by the labels 'I'.

Signal currents corresponding to the desired combinations of the black, red and green signals are extracted from the resistive plane by contacts disposed midway between the black, red and green injection points. These current extraction points are indicated by 'E' in Figure 38. This technique provides the same weight distributions for the black, red and green beams, in contrast with the previous example that had single weights of 1 for the black beams and three equal weights of 1/root(3) for the red and green beams. Extracted currents are fed to "virtual earth" amplifier inputs or low-input impedance amplifiers such as grounded base bipolar transistors. The set of weights realized by this technique can be tailored by choice of the shape and size of the current injection and extraction contact lands. No simple rule is proposed for deciding the size and shape - a proposition must simply be verified by carrying out a two-dimensional finite-element computer analysis of the current flow in and potentials existing on the resistive sheet.

Once the combined signals have been produced by the I.F. beam forming network they are fed to a bank of upconvertors 3730 for frequency

translation to the desired satellite-mobile frequency band. The upconvertors are all driven by the same local oscillator signal so as to preserve the relative phase of the signals, and have matched gain to preserve relative amplitudes. The upconverted signals may then be amplified by a matrix power amplifier 3740 to raise the power level to the desired transmit power.

The inventive technique described above can be extended to produce any number of virtual beams associated with subdivisions of the total frequency band available. In the three-color example, each "color" is associated with a 1/3rd sub-bandwidth. If 16.5MHz total is available, for example, each transponder channel bandwidth can be nominally 5.5 MHz. If the number of feeds is 37, for example, 37, 5.5MHz "black" beams are generated, 37, 5.5MHz wide red beams and 37, 5.5MHz green beams. Thus the total bandwidth available for communication is 37 times 16.5MHz, as it would have been had it been possible to employ immediate frequency re-use of all 16.5MHz in the "black" beams only. Thus the present invention provides the same efficient use of bandwidth as an immediate frequency re-use pattern but with a considerably improved C/I.

The extra capacity thus available is, in the FDMA version, obtained by increasing the number of transponder channels and thus hardware complexity proportionally. It will now be shown how an exemplary TDMA embodiment can be advantageously constructed in which the capacity increase is obtained without increased hardware complexity.

Figure 39 illustrates the exemplary TDMA embodiment. In this case the number of transponder channels 3910 is the same as the number of antenna feeds, and the bandwidth of each channel is the full bandwidth available to the system. The I.F. beam forming network 3920 also functions as previously described to synthesize black, red and green beams, but only one color is connected at a time to the set of transponder channels by virtue of the commutating switches 3911. Either (1) all transponder channels are connected to a corresponding number of "black" beam inputs, or, (2) by operating switches 3911 all at the same time, all transponders are connected

to the red beam inputs, or (3), as shown in Figure 39, to the green beam inputs.

The switches are cycled such that for a first portion of a TDMA frame period (e.g., 1/3rd) the black beams are used, for a second portion of the time the red beams are energized, and for a third portion of the time the green beams are energized. The time periods during which the switches dwell in each position do not have to be equal, and can in principle be adapted according to which color has the highest instantaneous capacity demand in any cell. The functioning of the rest of the transponder is as previously described for the FDMA case.

It will be appreciated that the commutation of switches 3911 is synchronized with transmissions from the central ground station or stations, and this can be achieved by any of a variety of techniques, such as providing an on-board clock that can be programmed form the ground to execute the regular cycle of switch operations and to synchronize the ground station transmissions to the satellite, which is the master timer. Alternatively a ground station can transmit a switch command using a control channel separate from the traffic channels. The method of achieving synchronism of the beam gyrations with the ground network is immaterial to the principle of the present invention.

It may be appreciated that, although both the TDMA and FDMA versions of the invention disclosed above used fixed beam forming networks, it is possible by an obvious extension of the method to permute the allocation of frequencies or time slots to beam colors by use of switches 3911 controlled from the ground in such a way as to keep the areas of the earth illuminated by a given frequency or time slot as nearly as possible fixed. This is of course achieved more accurately by using greater numbers of "colors" (that is timeslots or sub-bands). Increasing the number of sub-bands involves hardware complexity in the FDMA case, so the TDMA version is preferred in this respect. The phasing of the commutation switches may thus be chosen so as to compensate for satellite motion and

keep the areas illuminated by a particular timeslot or frequency more or less constant. The present invention can be applied to any number of timeslots and sub-bands, and in the latter case a digital implementation comprising analog-to-digital conversion of the transponder signals, digital filtering and digital beam forming using digital weight multiplication may be advantageous.

The above-described exemplary embodiments are intended to be illustrative in all respects, rather than restrictive, of the present invention. Thus the present invention is capable of many variations in detailed implementation that can be derived from the description contained herein by a person skilled in the art. All such variations and modifications are considered to be within the scope and spirit of the present invention as defined by the following claims.

WHAT IS CLAIMED IS:

1. A transmitting apparatus for transmitting a plurality of signals to a plurality of receivers using the same radio frequency channel comprising:

channel processing means for converting each of said signals to a numerical sample stream representative of a modulated radio wave;

matrix processing means for forming numerical combinations of said numerical sample streams using a set of matrix coefficients;

conversion means for converting said numerical combinations to corresponding analog modulated radio signals on a designated frequency and amplifying said modulated radio signals to a transmit power level;

antenna means, coupled to said conversion means, for transmitting said modulated radio signals; and

control means for adjusting said matrix coefficients such that each of said receivers receives an intended one of said plurality of signals with substantially zero interference from remaining, unintended signals.

2. A transmitting apparatus for transmitting a plurality of signals to a plurality of receivers using the same radio frequency channel comprising:

channel processing means for converting each of said signals to a numerical sample stream representative of a modulated radio wave;

matrix processing means for forming numerical combinations of said numerical sample streams using a set of matrix coefficients;

conversion means for converting said numerical combinations to corresponding analog modulated radio signals on a designated frequency and amplifying said modulated radio signals to a transmit power level;

antenna means, coupled to said conversion means, for transmitting said modulated radio signals;

control means for adjusting said matrix coefficients to maximize at a worst case receiver a received signal level of an intended one

of said plurality of signals relative to a received interference level of the remaining, unintended signals.

3. A receiving apparatus for receiving a plurality of signals from a plurality of transmitters using the same radio frequency channel and producing a plurality of output signals comprising:

multiple-element antenna means for receiving said transmitted signals and producing a received signal from each element;

multi-channel radio-frequency processing means for filtering, amplifying and converting each of said received signals from respective antenna elements into a corresponding numerical sample stream;

numerical matrix processing means for combining said numerical sample streams using a set of matrix coefficients to produce a plurality of separated signals;

channel processing means for processing each of said separated signals to produce said output signals; and

control means for adjusting said matrix coefficients such that each of said output signals corresponds to an intended one of said transmitted signals with substantially zero interference from the remaining, unintended transmitted signals.

4. A receiving apparatus for receiving a plurality of signals from a plurality of transmitters using the same radio frequency channel and producing a plurality of output signals comprising:

multiple-element antenna means for receiving said transmitted signals and producing a received signal from each element;

multi-channel radio-frequency processing means for filtering, amplifying and converting each of said received signals from respective antenna elements into a corresponding numerical sample stream;

numerical matrix processing means for combining said numerical sample streams using a set of matrix coefficients to produce a plurality of separated signals;

channel processing means for processing each of said separated signals to produce said output signals; and

control means for adjusting said matrix coefficients to maximize a correspondence between each of said output signals and a respectively intended one of said transmitted signals.

5. A transmitting apparatus for transmitting a plurality of signals to a corresponding plurality of receivers using a plurality of radio channels comprising:

grouping means for grouping selected ones of said signals into subsets for transmission using a same radio channel for each member of a subset and different radio channels for different subsets;

channel processing means for processing said signals into numerical signals representative of modulated radio signals;

matrix processing means, for each of said subsets, for forming numerical combinations of said signals belonging to the same subset using a set of matrix coefficients for each subset;

conversion means, for each subset, for converting said numerical combinations into corresponding analog modulated radio signals on the radio channel designated for each subset;

a plurality of combining means for combining one of said analog modulated signals from each subset to form an antenna element drive signal and amplifying said antenna element drive signal to a transmit power level;

multiple-element antenna means, wherein each of said antenna elements is coupled to a respective one of said combining means, for transmitting said antenna element drive signals; and

control means for controlling said grouping means and said matrix coefficients such that each of said receivers receives an intended signal at a desired signal strength with substantially zero interference from the remaining, unintended signals.

6. The transmitting apparatus of claim 5, wherein said control means is also for minimizing a sum of said transmit power levels.

7. A receiving apparatus for receiving a plurality of signals from a corresponding plurality of transmitters using a plurality of radio channel's frequencies and producing a plurality of output signals comprising:

multiple-element antenna means for receiving said plurality of transmitted signals and producing a composite received signal from each element;

multi-channel radio frequency processing means for filtering, amplifying, splitting and converting each of said composite received signals into a numerical sample streams, each numerical sample stream corresponding to a portion of a respective composite signal received on a corresponding radio frequency channel;

numerical matrix processing means for producing numerical combinations of said numerical sample streams corresponding to a same radio channel frequency using a set of matrix coefficients associated with each channel frequency;

channel processing means for processing said numerical combinations to produce said plurality of output signals; and

control means for controlling said matrix coefficients such that said output signals each correspond to an intended one of said transmitted signals with substantially zero interference from the remaining, unintended signals.

8. A transmitting apparatus for transmitting a plurality of signals to a corresponding plurality of receivers by dividing a transmission time on a same radio frequency channel into a plurality of timeslots comprising:

grouping means for grouping selected ones of said signals into subsets for transmission in a same designated timeslot on said radio channel, whereby signals in a same subset use a same timeslot and signals in different subsets use different timeslots;

channel processing means for processing said signals for transmission in their designated timeslots;

multiplexing means for multiplexing one signal from each subset into its respective timeslot to form Time Division Multiplex signals;

matrix processing means for producing numerical combinations of said Time Division multiplex signals using a different set of matrix coefficients for each timeslot;

a plurality of conversion means for converting said numerical combinations to corresponding modulated analog radio signals on said radio channel and amplifying them to a transmit power level;

multiple-element antenna means, wherein each of said antenna elements is coupled respectively to one of said conversion means, for transmitting said analog radio signals; and

control means for controlling said grouping means and said matrix coefficients such that each of said plurality of receivers receives an intended signal at a desired signal strength with substantially zero interference from the remaining, unintended signals.

9. The transmitting apparatus of claim 8, wherein said control means is also for minimizing a sum of said transmit power levels.

10. A receiving apparatus for receiving a plurality of signals from a corresponding plurality of transmitters on a same radio channel frequency

using a plurality of timeslots in a Time Division multiple Access frame period and producing a plurality of output signals, comprising:

multiple-element antenna means for receiving said plurality of transmitted signals and producing a composite received signal from each element;

multi-channel radio frequency processing means for filtering, amplifying, and converting each of said composite received signals into a corresponding numerical sample stream;

numerical matrix processing means for producing numerical combinations of said numerical samples streams using a different set of matrix coefficients for each of said TDMA timeslots;

demultiplexing means to divide each of said numerical combinations into said plurality of timeslots to produce decimated signal streams;

channel processing means for processing said decimated signal streams to produce said plurality of output signals; and

control means for controlling said matrix coefficients such that said output signals each correspond to an intended one of said transmitted signals with substantially zero interference from the remaining, unintended signals.

11. A transmitting apparatus for transmitting a plurality of signals to a corresponding plurality of receivers using a plurality of radio frequency channels and by dividing the transmission time on each radio frequency channel into a plurality of timeslots comprising:

grouping means for grouping selected ones of said signals into subsets for transmission in a same timeslot and frequency channel whereby signals in a same subset use a same timeslot and frequency and signals in different subsets use different frequencies or timeslots;

channel processing means for processing said signals transmission in their designated timeslots and on their designated frequencies;

multiplexing means for multiplexing one signal from each subset using the same frequency into its respective timeslot to form a plurality of Time Division Multiplex signals for transmission on each of said frequency channels;

matrix processing means, associated with each radio frequency channel, for producing numerical combinations of said Time Division Multiplex signals using a different set of matrix coefficients for each timeslot and frequency channel;

conversion means, associated with each radio frequency channel, for converting said numerical combinations to corresponding modulated analog radio signals;

a plurality of combining means for combining one of said analog radio signals from each frequency channel and amplifying the combined signal to a transmit power level;

multiple-element antenna means, wherein each of said antenna elements is coupled to a respective one of said combining means, to transmit said amplified, combined signals; and

control means for controlling said grouping means and said matrix coefficients such that each of said receivers receives an intended signal at a desired signal strength with substantially zero interference from the remaining, unintended signals.

12. The transmitting apparatus of claim 11 wherein said control means is also for minimizing a sum of the transmit power levels.

13. A receiving apparatus for receiving a plurality of signals from a corresponding plurality of transmitters using a plurality of timeslots in a

Time Division Multiple Access frame period and a plurality of radio channel frequencies and producing a plurality of output signals, comprising:

multiple-element antenna means for receiving said plurality of transmitted signals and producing a composite received signal from each element;

multi-channel radio frequency processing means for filtering, amplifying, splitting and converting each of said composite received signals into a number of numerical sample streams corresponding to said plurality of radio channel frequencies;

numerical matrix processing means for producing numerical combinations of said numerical samples streams associated with a same radio channel using sets of matrix coefficients associated with each timeslot and frequency combination;

demultiplexing means for dividing each of said numerical combinations into said plurality of timeslots to produce decimated signal streams;

channel processing means for processing said decimated signal streams to produce said plurality of output signals; and

control means for controlling said matrix coefficients such that said output signals each correspond to an intended one of said transmitted signals with substantially zero interference from the remaining, unintended signals.

14. A method of wireless communication with mobile stations comprising the steps of:

transmitting signals representative of voice or data communications from said mobile stations to an orbiting satellite having a multi-element antenna;

receiving combinations of said signals at each of said antenna elements and coherently transponding said signal combinations to a ground station;

receiving said coherently transponded signals at said ground station and analog-to-digital converting said coherently transponded signals to produce corresponding numerical sample streams;

processing said numerical sample streams in a numerical matrix processor to separate said transmitted signals originating respectively from each of said mobile stations to produce separated sample streams; and

numerically processing said separated sample streams to reconstruct said voice- or data-representative signals and sending said reconstructed signals to a telephone switching network.

15. A method of wireless communication with mobile stations comprising the steps of:

receiving, from a telephone switching network, signals representing voice or data destined for each of said mobile stations;

digitizing said voice- or data-representative signals and processing said digitized signals into corresponding streams of digital samples representing modulated signals;

combining said modulated-signal-representative sample streams using a numerical matrix processor to produce sample streams representative of antenna element signals;

digital-to-analog converting said antenna element signals to modulate corresponding ground station transmission means and transmitting said modulated signals to an orbiting satellite in such a way as to preserve their relative phase and amplitude relationships; and

receiving said modulated signals at said orbiting satellite from said ground station and transponding each using a corresponding antenna element such that each of said data- or voice-representative signals is transmitted to its intended mobile station destination and not to the other mobile stations for which it is not intended.

16. Apparatus for wireless communication with mobile stations comprising:

digitizing means for digitizing voice- or data-representative signals received from a telephone switching network if not already in digital form and processing said signals into corresponding streams of digital samples representing modulated signals;

numerical matrix processor means for combining said modulated-signal-representative sample streams in order to produce sample streams representative of antenna element signals;

digital-to-analog conversion means to convert said antenna element signals into corresponding transmitter modulating waveforms;

ground station transmitter means each modulated by a corresponding one of said modulation waveforms for transmitting said waveforms to an orbiting satellite in such a way as to preserve their phase and amplitude relationships;

satellite receiving means in said orbiting satellite for receiving and demodulating each of said modulating waveforms; and

satellite transponder means for converting each of said demodulated waveforms to a new frequency, amplifying them to a transmit power level and transmitting each through a corresponding element of a multiple-element satellite antenna such that each of said data- or voice-representative signals is transmitted to its intended destination mobile station and not to the other mobile stations for which it is not intended.

17. An apparatus for minimizing co-channel interference in radiocommunication systems comprising:

transceiver means for receiving or transmitting signals;

signal processing means for converting each of said signals into a numerical sample stream;

matrix processing means for producing numerical combinations of said numerical sample streams using a set of matrix coefficients, and

control means for adjusting said matrix coefficients to minimize co-channel interference between each of said signals.

18. The apparatus of claim 17, wherein said matrix processing means further comprises:

matrix multiplication means for multiplying an inverse of a matrix having said matrix coefficients with a matrix having coefficients related to sampled beams.

19. The apparatus of claim 18, wherein said matrix processing means further comprises:

means for inverting a matrix having said matrix coefficients every sample period.

20. The apparatus of claim 18, wherein said matrix processing means further comprises:

means for inverting a matrix having said matrix coefficients at when a new call is connected.

21. The apparatus of claim 17, wherein said control means further comprises:

means for preventing a matrix having said matrix coefficients from becoming numerically ill-conditioned.

22. The apparatus of claim 21, wherein said preventing means further comprises:

means for changing a radio channel of a mobile unit when at least two mobile units approach a common location.

23. A method for transmitting signals to a plurality of remote units comprising the steps of:

correlating a signal received from a new remote unit during random access transmission with a plurality of individual antenna beam element signals to determine a new column of coefficients for a receive matrix;

determining a new inverse C-matrix for receiving traffic from the new remote unit based on an old inverse C-matrix and the new column;

transforming the new column to a new transmit C-matrix row by scaling relative coefficient phase angles using a ratio of up- to down-link frequencies;

determining a new transmit inverse C-matrix based on the old transmit inverse C-matrix and the new transmit C-matrix row, and

transmitting signals to said remote units using said new transmit inverse C-matrix.

24. The method of claim 23 further comprising the step of:

repeating each of the steps set forth therein when a new mobile unit makes a random access transmission.

25. An apparatus for transmitting signals to a plurality of remote units comprising:

means for correlating a signal received from a new remote unit during random access transmission with a plurality of individual antenna beam element signals to determine a new column of coefficients for a receive matrix;

means for determining a new inverse C-matrix for receiving traffic from the new remote unit based on an old inverse C-matrix and the new column;

means for transforming the new column to a new transmit C-matrix row by scaling relative coefficient phase angles using a ratio of up- to down-link frequencies;

means for determining a new transmit inverse C-matrix based on the old transmit inverse C-matrix and the new transmit C-matrix row, and

means for transmitting signals to said remote units using said new transmit inverse C-matrix.

26. A method of allocating a communications channel for transmitting one of a first plurality of signals to one of a corresponding plurality of receivers using a second plurality of transmitter-antenna element combination including:

estimating a coefficient relating to phase and amplitude of propagation of said signal from each of said transmitter-antenna element combinations to said receiver;

processing said coefficients with similar coefficients estimated for groups of other of said plurality of receivers that use the same communications channel within their respective groups and different channels for different groups in order to determine a figure of merit for each group;

allocating to said signal the communications channel used by the group producing the highest figure of merit in the above step.

27. A method of allocating a communications channel according to claim 26 in which said figure of merit for a group is related to the total transmitter power of said transmitter-antenna element combinations if said signal is allocated the same communications channel as said group.

28. The method of claim 26 in which said communications channel is one of a number of timeslots of a time-division-multiplex transmission frame each timeslot being used by one of said groups of receivers.

29. The method of claim 26 in which said communications channel is one of a number of radio frequency channels used by one of said groups of receivers.

30. The method of claim 26 in which said communications channel is one of a number of combinations of radio frequency channel and TDM timeslot each combination being used by one of said groups of receivers.

31. A method for communicating a plurality of first signals from a first station to a corresponding plurality of second stations and a plurality of second signals from said second stations to said first station involving:

processing said first signals into a form representative of modulated radio signals;

combining said representations to form a number of transmit signals using a set of combining parameter;

transmitting said transmit signals using a corresponding transmitter and antenna;

measuring at said second stations an amount received of at least one unwanted signal relative to the amount of wanted signal received;

coding said measurements into said second signals and transmitting them from said second stations to said first stations;

receiving said coded measurements and decoding them at said first station and using said results to modify said combining parameters such that unwanted signals received at said second stations are reduced and wanted signals maximized.

32. A method for communicating a plurality of first signals from a first station to a corresponding plurality of second stations and a plurality of second signals from said second stations to said first station involving:

processing said first signals into a form representative of modulated radio signals;

combining said representations to form a number of transmit signals using a set of combining parameter;

amplifying said transmit signals to desired transmit power levels and transmitting them using a corresponding transmitter and antenna for each;

measuring at said second stations an amount received of at least one unwanted signal relative to the amount of wanted signal received;

coding said measurements into said second signals and transmitting them from said second stations to said first station;

receiving said coded measurements and decoding them at said first station and using said results to modify said combining parameters such that unwanted signals received at said second stations are reduced and the sum of said transmit power levels needed to transmit said first signals is minimized.

33. A method according to claim 31 or claim 32 in which said first station comprises a ground station communicating with at least one orbiting satellite.

34. A method according to claim 33 in which said corresponding transmitters form part of a satellite transponder for relating said transmit signals from said ground station via said satellite.

35. A method according to claim 34 in which said transmit signals are relayed by a coherent transponder in such a way as to preserve their relative phases and amplitudes.

36. A coherent satellite transponder for transponding a plurality of signals received from a ground station comprising:

a corresponding plurality of receiving means for receiving said signals using frequency or phase modulation on a corresponding number of carrier frequencies and demodulating them to produce video signals;

upconverting means for translating each of said video signals to a new carrier frequency;

transmitter-antenna means for amplifying each of said translated signals to a desired transmit power level and transmitting them using a corresponding antenna means.

37. A coherent satellite transponder for transponding signals received form a plurality of ground stations comprising:

receiver-antenna means for receiving combinations of said signals from said ground stations;

downconverting means for translating each of said received combinations to a corresponding video signal or signals;

transmitter-modulator means for modulating each of said video signals onto a corresponding carrier frequency using frequency or phase modulation, amplifying them to desired transmit power levels and transmitting them using transmit antenna means.

38. A transponder according to claim 35 in which approximately half of said plurality of receiving means use Right Hand Circular antenna polarization and the rest use Left Hand Circular polarization.

39. A transponder according to claim 36 in which approximately half of said transmitter-modulator means transmit using Right Hand Circular antenna polarization and the rest use Left Hand Circular polarization.

40. A transponder according to claim 36 in which said corresponding carrier frequencies for said RHC polarized transmissions are the same as those for LHC transmissions.

41. A transponder according to claim 36 in which said corresponding carrier frequencies for said RHC polarized transmissions lie interleaved between those of LHC transmissions.

42. A transponder according to claim 37 in which said corresponding carrier frequencies used by receiving means using RHC polarization are the same as those of receiving means using LHC polarization.

43. A transponder according to claim 37 in which said corresponding carrier frequencies used by receiving means using RHC polarization are interleaved between those of receiving means using LHC polarization.

44. A satellite communications system employing a multiple element antenna receiving signals on a first frequency and relaying them to a ground station on a second frequency band including:

downconverting means for converting signals received at each of said multiple antenna elements on said first frequency to a corresponding baseband signal;

multiplexing means, for time-division multiplexing said corresponding baseband signals to form a multiplexed sample stream;

modulator means for modulating a carrier in said second frequency band with said multiplexed sample stream and transmitting said modulated carrier to said ground station.

45. The system as in claim 44, in which said downconverting means are quadrature downconverting means producing an I and a Q baseband signal.

46. The system as in claim 44, in which said multiplexing means comprises an I-signal multiplexing means and a Q-signal multiplexing means.

47. The system as in claim 44, in which said modulator means is a quadrature modulator means in which an I signal and a Q signal are impressed on respective carrier signals of nominally 90 degrees phase difference.

48. The system of claim 45, in which said multiplexing means comprises analog to digital conversion of said I and Q baseband signals and digital multiplexing of the resultant digital streams to form I and Q bitstreams.

49. The system of claim 48, in which said modulator means is a digital modulator means in which said multiplexed I and Q bit streams digitally modulate respective quadrature carriers at said second frequency.

50. The system of claim 44, in which said corresponding baseband signals are representative of the instantaneous phase and amplitude respectively of said antenna signals relative to common phase and amplitude references.

51. A satellite communications system for relaying signals received from a ground station on a second frequency band by transmitting them on a first frequency using a multiplicity of antenna elements, including:

downconverting means for converting signals received at said second frequency to a corresponding baseband signal;

demultiplexing means, for time-division demultiplexing said corresponding baseband signal to obtain separate sample streams;

separate modulator means corresponding to each of said separate sample streams for modulating a carrier at said first frequency with said separate sample streams to generate corresponding modulated signals;

transmit amplifier means for amplifying said modulated signals and transmitting them using said multiple antenna elements.

52. The system as in claim 51, in which said downconverting means is a quadrature downconverting means producing an I and a Q baseband signal.

53. The system as in claim 51, in which said demultiplexing means comprises an I-signal demultiplexing means and a Q-signal demultiplexing means and said separate sample streams comprise I and Q sample streams.

54. The system as in claim 53, in which said separate modulator means include quadrature modulator means in which an I sample stream and a Q sample stream are impressed on respective carrier signals having nominally 90 degrees phase difference.

55. The system of claim 52, in which said demultiplexing means is a digital demultiplexing means producing separate I and Q bitstreams.

56. The system of claim 55, in which said separate I and Q bitstreams are analog to digital converted and filtered to form corresponding I and Q analog waveforms.

57. The system of claim 56, in which said I and Q waveforms each modulate a carrier at said first frequency using respective quadrature modulators to generate said corresponding modulated signals.

58. The system of claim 51, in which said corresponding baseband signal includes an amplitude-corresponding signal and a phase-corresponding signal.

59. The system of claim 51, in which said separate modulator means comprise an amplitude modulator and a phase modulator.

60. A bidirectional satellite communications system comprising:

a satellite including:

a multiple element antenna having multiple antenna elements for receiving signals on a first frequency and relaying them to a ground station on a second frequency band;

downconverting means for converting signals received at each of said multiple antenna elements on said first frequency to a corresponding baseband signal;

multiplexing means, for time-division multiplexing said corresponding baseband signals to form a multiplexed sample stream; and

modulator means for modulating a carrier in said second frequency band with said multiplexed sample stream and transmitting said modulated carrier to said ground station and

a ground station including:

downconverting means for converting signals received at said second frequency to a corresponding baseband signal;

demultiplexing means, for time-division demultiplexing said corresponding baseband signal to obtain separate sample streams;

separate modulator means corresponding to each of said separate sample streams for modulating a carrier at said first frequency with said separate sample streams to generate corresponding modulated signals;

transmit amplifier means for amplifying said modulated signals and transmitting them to said satellite using multiple antenna elements; and

wherein said multiplexing means and said demultiplexing means are synchronized on board the satellite to the same clock.

61. The system of claim 60, in which said ground station further comprises:

means for synchronizing said demultiplexing means to a time multiplexed signal received from said satellite on said second frequency band and synchronizing a second multiplexing means for transmission of a time multiplexed signal to said satellite based on propagation delay so that said transmitted signal will arrive in synchronism with a second demultiplexing means on board said satellite.

62. The system of claim 61, in which said ground station synchronizes said demultiplexing means using pilot samples transmitted by said satellite in the time-multiplexed signal.

63. The system of claim 44, in which said ground station synchronizes said demultiplexing means using pilot samples transmitted by said satellite in the time-multiplexed signal.

64. The system according to claim 63, in which said pilot symbols include the null symbol (0,0).

65. The system according to claim 63, in which said pilot symbols are used at said ground station to assist in synchronizing a demultiplexer.

66. The system according to claim 63, in which said pilot symbols are used at said ground station to correct for transmission or modulation errors.

67. The system according to claim 44, in which said ground station includes an equalizer for reducing intersample interference arising in the transmission of said multiplexed sample stream.

68. The system of claim 51, in which said ground station includes a pre-equalizer to reduce intersample interference between said separate sample streams arising in the transmission from said ground station to said satellite in said second frequency band.

69. A system for communicating between one or more central ground stations and a plurality of mobile or portable stations via at least one orbiting satellite comprising:

at least one central ground station having transmitter means to transmit a plurality of signals to a satellite for transponding;

at least one satellite having multi-channel transponder means for receiving said plurality of signals, frequency translating, filtering and amplifying them and forming weighted combinations of them for applying after power amplification to a multiple element antenna such that a first subset of said signals will be radiated using a first set of antenna beams having radiation patterns that touch at approximately the -4dB down point relative to beam peak and such that at least one other subset of said signals will be radiated by another set of antenna beams displaced from said first set of beams by less than the -4dB beam diameter.

70. A system for communicating between one or more central ground stations and a plurality of mobile or portable stations via at least one orbiting satellite comprising:

at least one central ground station having transmitter means to transmit a plurality of signals to a satellite for transponding;

at least one satellite having multi-channel transponder means for receiving said plurality of signals, frequency translating, filtering and amplifying them and sequentially selecting weighted combinations of them for applying after power amplification to a multiple element antenna such that said signals are radiated for a first proportion of a time period using a first set of antenna beams having radiation patterns that touch at approximately the -4dB down point relative to beam peak and such that for at least one other proportion of said time period said signals are radiated by at least one other set of antenna beams displaced from said first set of beams by less than the -4dB beam diameter.

71. A method of communicating between a fixed ground station and a plurality of mobile stations using an orbiting satellite involving:

at said ground station grouping said mobile stations into a first number of sets containing not more than a second number of mobile stations that are each within their respective set mutually separated by a given minimum distance on the ground;

allocating the same communications channel to each of the mobiles in the same set, said communications channel comprising a frequency or a timeslot or a code or any unique combination of these;

transmitting signals from said ground station to said satellite for transponding to said mobile stations using their respectively allocated communications channels such that mobile stations in a first of said sets receive signals via a first set of satellite antenna beams that cross at approximately the -4dB down points relative to beam peak and a different set of mobile stations receive their intended signals via a different set of satellite antenna beams that are displaced from said first set of beams by less than said -4dB beam diameter.

72. A system for communicating between one or more central ground stations and a plurality of mobile or portable stations via at least one orbiting satellite comprising:

at least one central ground station having receiver means to receive a plurality of signals from said mobile stations transponded by said at least one satellite;

at least one satellite having multi-channel transponder means for receiving signals from said plurality of mobile stations using a multiple-element antenna and for frequency translating, filtering and amplifying signals received at said antenna elements and forming weighted combinations of them for applying to separate transmitter means for transmission to said ground station receiving means such that a first subset of said transmitted signals corresponds to a first set of mobile signals received in a first set of antenna beams having radiation patterns that touch at approximately the -4dB down point relative to beam peak and such that at least one other subset of said transmitted signals corresponds to mobile signals received by another set of antenna beams displaced from said first set of beams by less than the -4dB beam diameter.

73. A system for communicating between one or more central ground stations and a plurality of mobile or portable stations via at least one orbiting satellite comprising:

at least one central ground station having receiver means to receive a plurality of signals from said mobile stations transponded by said at least one satellite;

at least one satellite having multi-channel transponder means for receiving signals from said plurality of mobile stations using a multiple-element antenna and for frequency translating, filtering and amplifying signals received at said antenna elements and sequentially forming groups of differently weighted combinations of said signals that are applied sequentially to a separate transmitter means for each member of the same group for

transmission to said ground station receiving means such that a first group of said transmitted signals corresponds for a first proportion of a time period to a first set of mobile signals received in a first set of antenna beams having radiation patterns that touch at approximately the -4dB down point relative to beam peak and such that at least one other subset of said transmitted signals corresponds for a second proportion of said time period to mobile signals received by another set of antenna beams displaced from said first set of beams by less than the -4dB beam diameter.

74. A method of communicating between a fixed ground station and a plurality of mobile stations using an orbiting satellite involving:

at said ground station grouping said mobile stations into a first number of sets containing not more than a second number of mobile stations that are each within their respective set mutually separated by a given minimum distance on the ground;

allocating the same communications channel to each of the mobiles in the same set, said communications channel comprising a frequency or a timeslot or a code or any unique combination of these;

receiving signals at said ground station transponded by said satellite from said mobile stations using their respectively allocated communications channels such that mobile stations in a first of said sets are received by said satellite via a first set of satellite antenna beams that cross at approximately their -4dB down points relative to beam peak and a different set of mobile stations are received via a different set of satellite antenna beams that are displaced from said first set of beams by less than said -4dB beam diameter.

75. A system for wireless communications between a base station and a plurality of mobile stations comprising:

base station antenna means capable of generating transmit and receive antenna patterns in defined directions;

signal processing means coupled to said antenna means for defining said directions to be in association with particular communications channels;

channel allocation means for adaptively determining the communications channel to use for communicating with each of said mobile stations such as to optimize a figure of merit for said communications.

76. A system for communications between a fixed network and a plurality of mobile stations via an orbiting satellite comprising:

multi-beam satellite relay means capable of relaying different signals between mobile stations located in different regions covered by said beams and at least one earth station;

at least one earth station transmitting and receiving said relayed signals;

signal processing means for processing said relayed signals using a set of matrix coefficients such as to define the center of each of said regions to be associated with a particular communications channel;

control processor means for generating said matrix coefficients as a function of time using a set of satellite orbital parameters so as to maintain said regions associated with particular communications channels in fixed positions;

channel allocation means for adaptively allocating the communications channels for communications with each of said mobile stations such as to optimize a figure of merit for said communications.

77. A method of communication between a base station having a sectorized directional antenna and a plurality of mobile stations using TDMA involving:

rotating the sector radiation patterns of said antenna synchronously with the TDMA frame rate;

rotating the allocation of Digital Voice Color codes to sectors in the reverse direction at the same rate so that the same color code continues to be used for the same absolute direction.

78. A method according to claim 77 in which said antenna pattern rotation is accomplished by physically spinning the antenna.

79. A method according to claim 77 in which said antenna pattern rotation is accomplished electronically by means of phased array antenna signal processing.

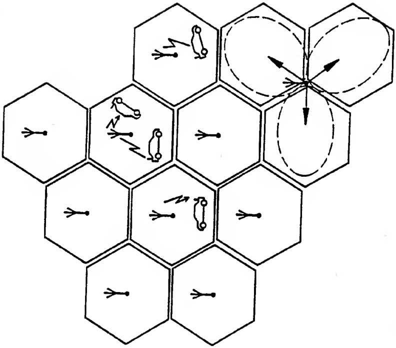


FIG. 1
(PRIOR ART)

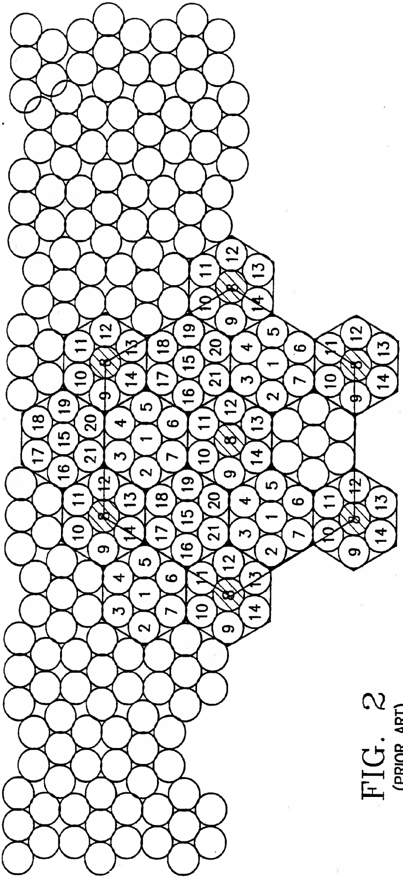


FIG. 2
(PRIOR ART)

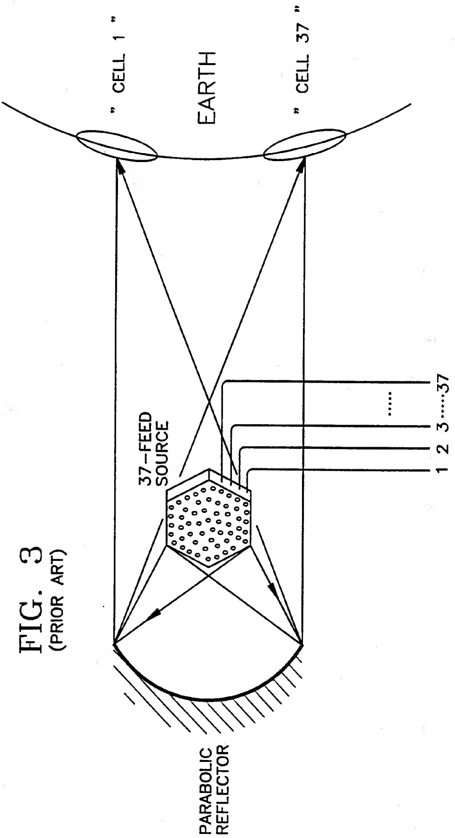


FIG. 3
(PRIOR ART)

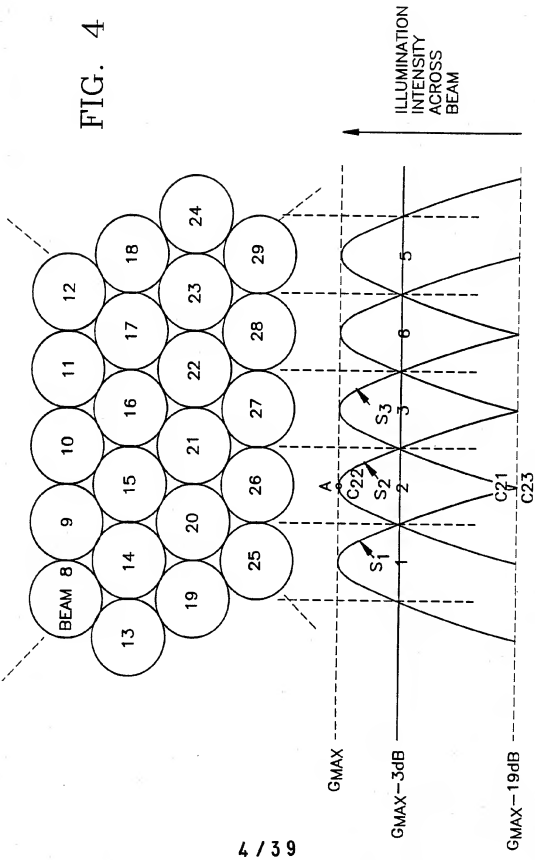

ILLUMINATION INTENSITY ACROSS BEAM
BEAM 8
9
10
11
12
13
14
15
16
17
18
19
20
21
22
23
24
25
26
27
28
29
GMAX
GMAX−3dB
GMAX−19dB
A
C22
C21
C23
S1
S2
S3
1
2
3
4
5

FIG. 4

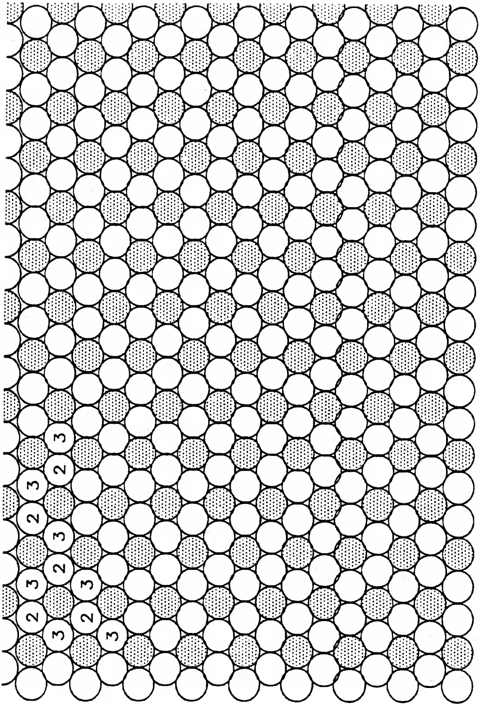


FIG. 5

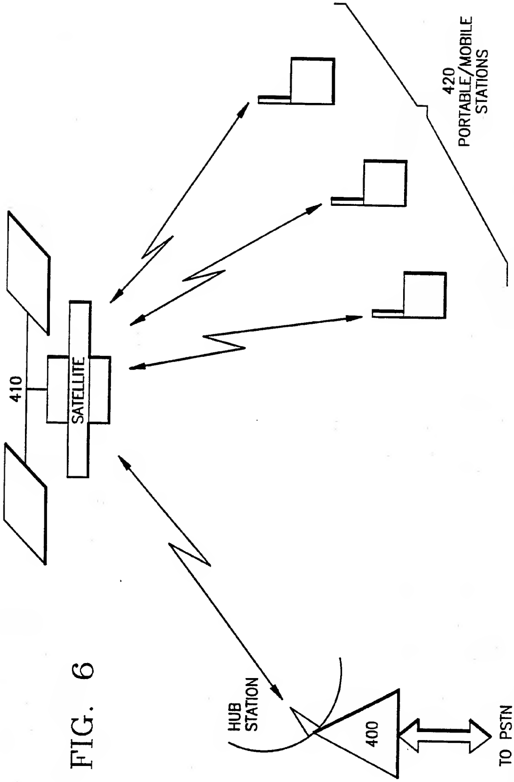

SATELLITE
410
420
PORTABLE/MOBILE
STATIONS
HUB
STATION
400
TO PSTN

FIG. 6

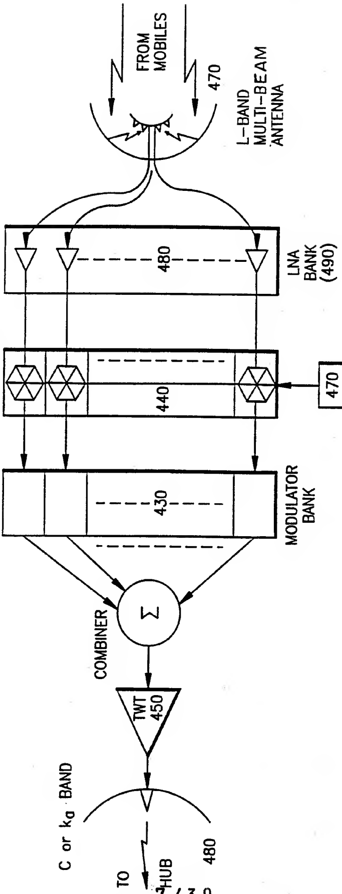

FROM MOBILES
470
L-BAND MULTI-BEAM ANTENNA
480
LNA BANK (490)
DOWNCONVERTER BANK
440
470 LO
430
MODULATOR BANK
Σ
COMBINER
TWT 450
C or k_a BAND
TO HUB
480

FIG. 7

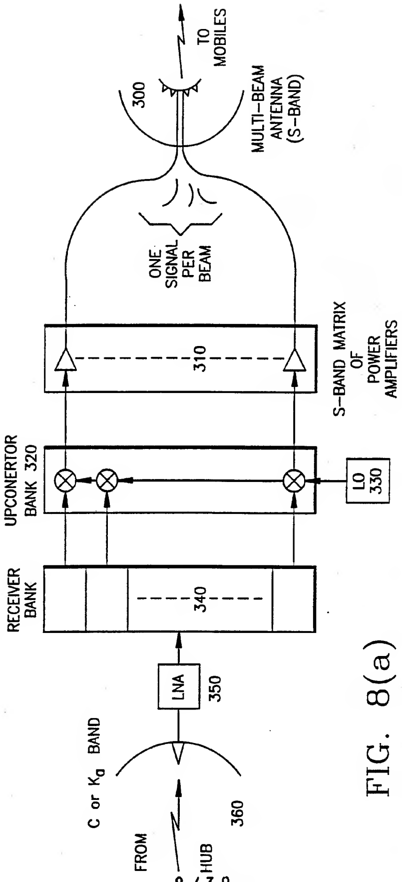

TO MOBILES
300
MULTI-BEAM ANTENNA (S-BAND)
ONE SIGNAL PER BEAM
310
S-BAND MATRIX OF POWER AMPLIFIERS
UPCONERTOR BANK 320
LO 330
RECEIVER BANK
340
LNA
350
C or K_a BAND
360
FROM HUB

FIG. 8(a)

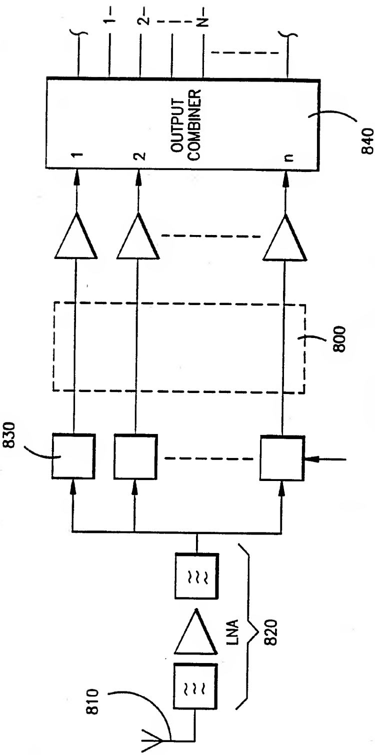


FIG. 8(b)

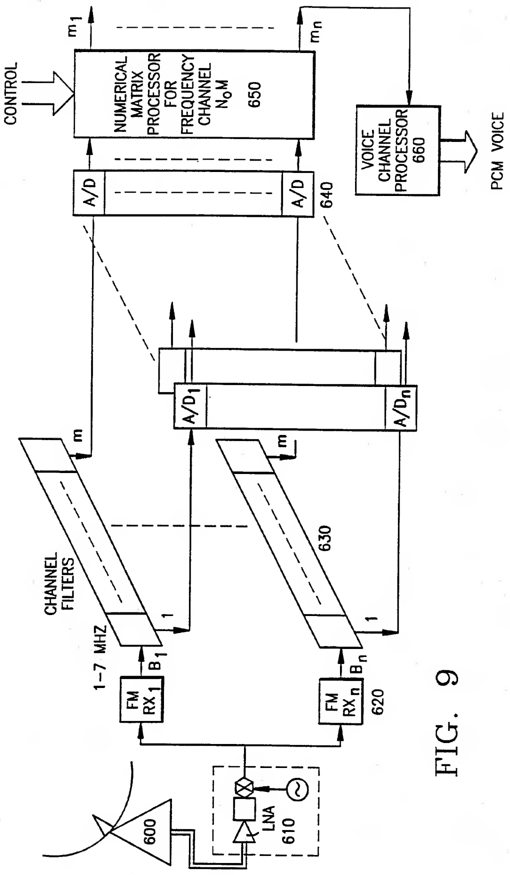
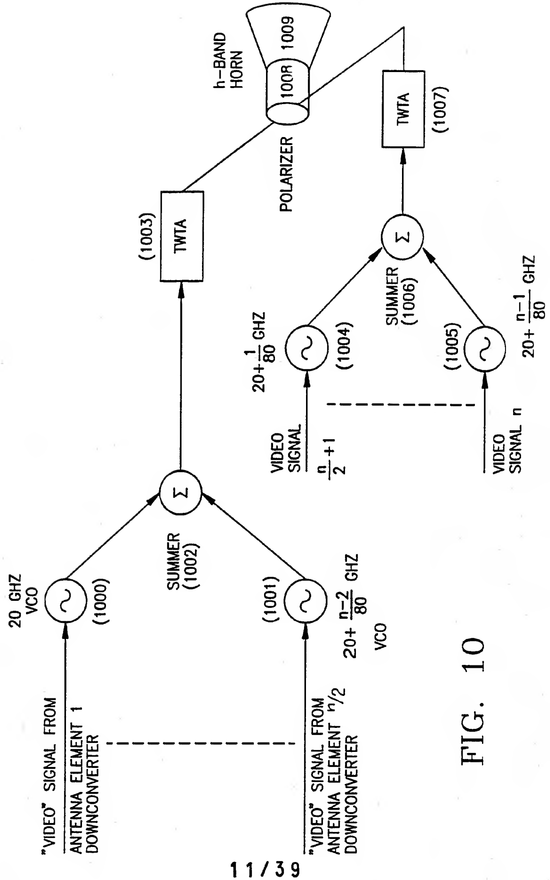

CONTROL
NUMERICAL MATRIX PROCESSOR FOR FREQUENCY CHANNEL NoM
650
m1
mn
A/D
A/D
640
VOICE CHANNEL PROCESSOR 660
PCM VOICE
A/D1
A/Dn
m
m
CHANNEL FILTERS
630
1
1
1-7 MHZ
B1
Bn
FM RX1
FM RXn
620
600
LNA
610

FIG. 9



"VIDEO" SIGNAL FROM ANTENNA ELEMENT 1 DOWNCONVERTER
"VIDEO" SIGNAL FROM ANTENNA ELEMENT n/2 DOWNCONVERTER
20 GHZ VCO
(1000)
(1001)
20+ (n-2)/80 GHZ
VCO
SUMMER (1002)
(1003)
TWTA
POLARIZER
h-BAND HORN
1008
1009
TWTA
(1007)
SUMMER (1006)
20+ 1/80 GHZ
(1004)
(1005)
20+ (n-1)/80 GHZ
VIDEO SIGNAL n/2 +1
VIDEO SIGNAL n

FIG. 10

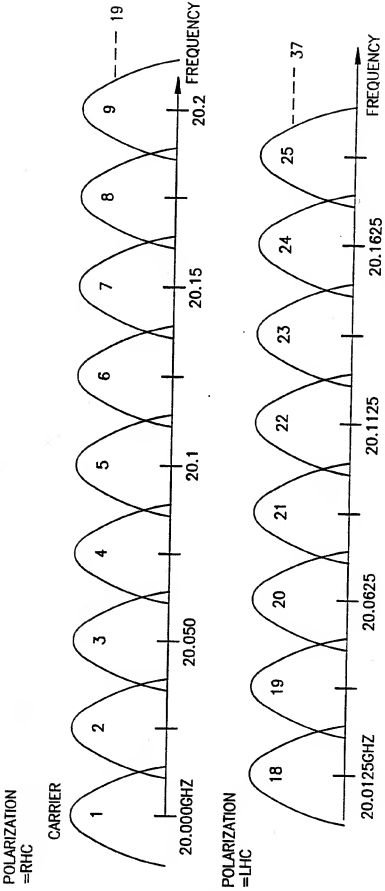

POLARIZATION =RHC
CARRIER
1
2
3
4
5
6
7
8
9
19
20.000GHZ
20.050
20.1
20.15
20.2
FREQUENCY
POLARIZATION =LHC
18
19
20
21
22
23
24
25
37
20.0125GHZ
20.0625
20.1125
20.1625
FREQUENCY

FIG. 11

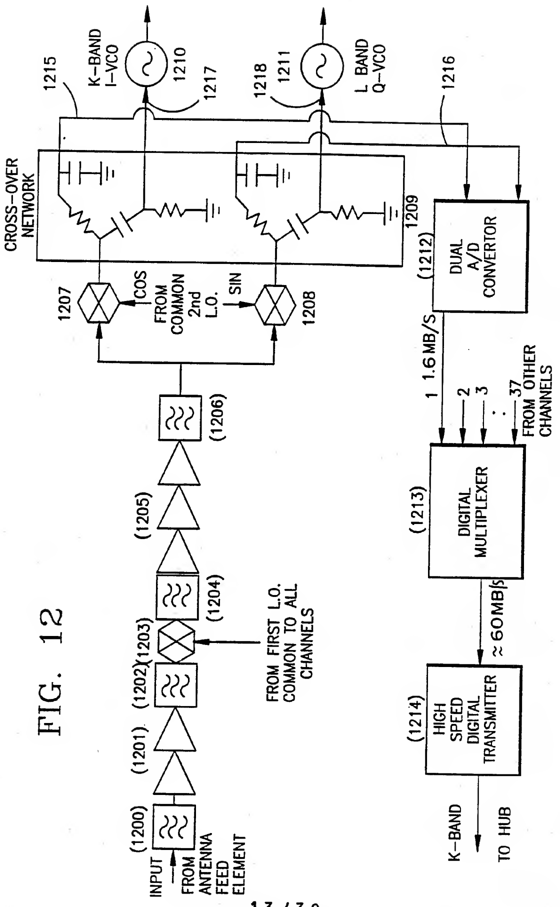

INPUT
FROM ANTENNA FEED ELEMENT
(1200)
(1201)
(1202)
(1203)
(1204)
(1205)
(1206)
FROM FIRST L.O. COMMON TO ALL CHANNELS
1207
1208
COS
FROM COMMON 2nd L.O.
SIN
CROSS-OVER NETWORK
1209
1215
K-BAND I-VCO
1210
1217
1218
1211
L BAND Q-VCO
1216
(1212)
DUAL A/D CONVERTOR
1 1.6 MB/S
2
3
37
FROM OTHER CHANNELS
(1213)
DIGITAL MULTIPLEXER
≈ 60MB/S
(1214)
HIGH SPEED DIGITAL TRANSMITTER
K-BAND
TO HUB

FIG. 12

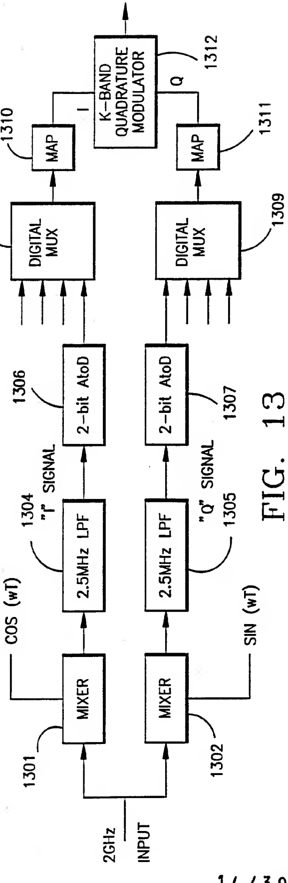

2GHz
INPUT
MIXER
MIXER
1301
1302
COS (wT)
SIN (wT)
2.5MHz LPF
2.5MHz LPF
1304
"I" SIGNAL
"Q" SIGNAL
1305
2-bit AtoD
2-bit AtoD
1306
1307
DIGITAL MUX
DIGITAL MUX
1308
1309
MAP
MAP
1310
1311
I
Q
K-BAND QUADRATURE MODULATOR
1312

FIG. 13

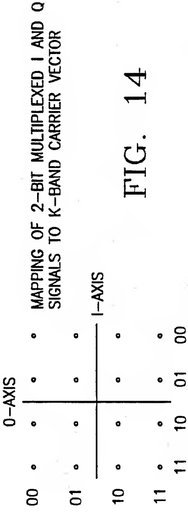

MAPPING OF 2-BIT MULTIPLEXED I AND Q SIGNALS TO K-BAND CARRIER VECTOR
Q-AXIS
I-AXIS
00
01
10
11
11
10
01
00

FIG. 14

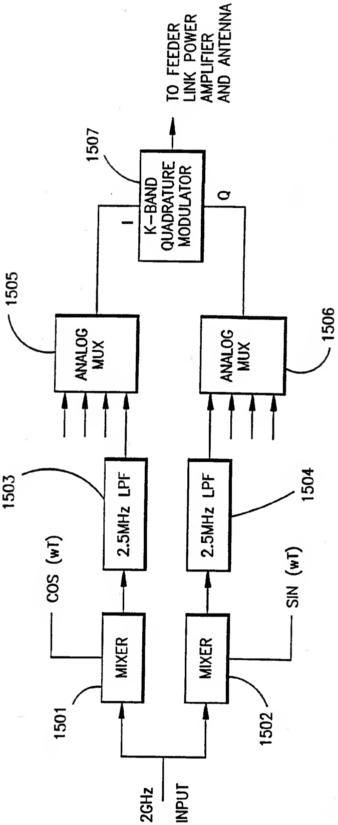

2GHz
INPUT
1501
1502
MIXER
MIXER
COS (wT)
SIN (wT)
1503
1504
2.5MHz LPF
2.5MHz LPF
1505
1506
ANALOG
MUX
ANALOG
MUX
I
Q
K-BAND
QUADRATURE
MODULATOR
1507
TO FEEDER
LINK POWER
AMPLIFIER
AND ANTENNA

FIG. 15

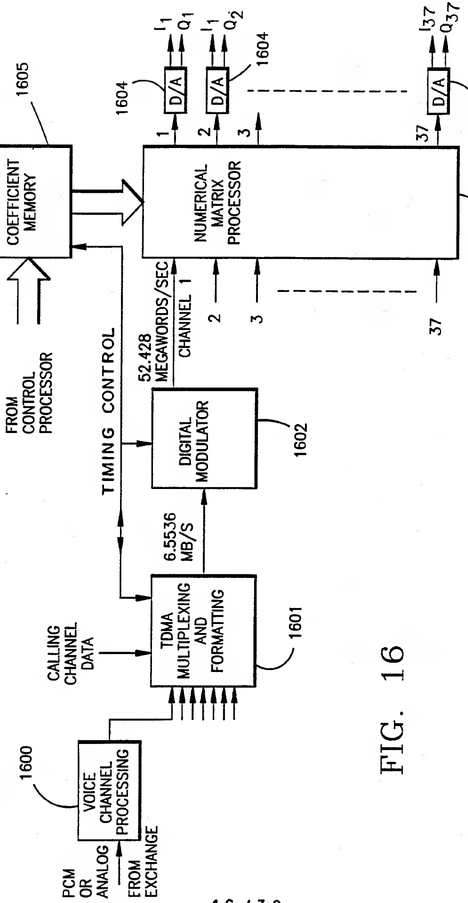


FIG. 16

SUBSTITUTE SHEET (RULE 26)

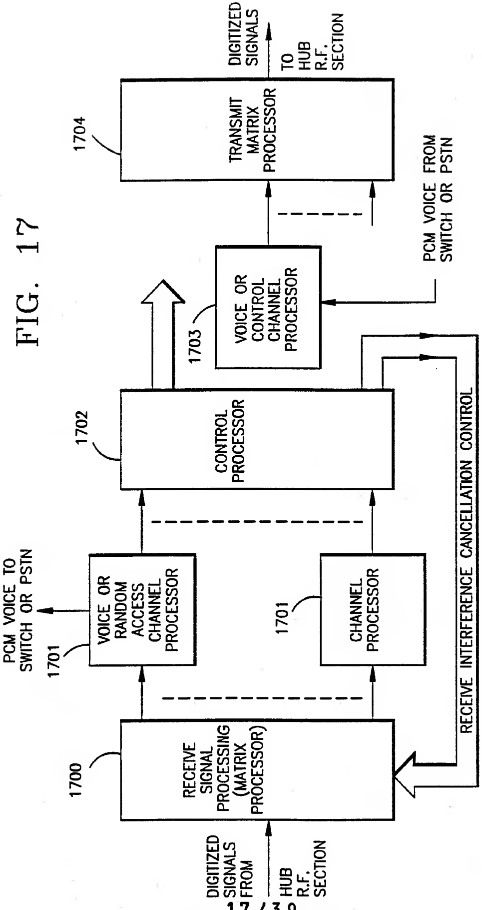

DIGITIZED SIGNALS FROM HUB R.F. SECTION
1700
RECEIVE SIGNAL PROCESSING (MATRIX PROCESSOR)
1701
VOICE OR RANDOM ACCESS CHANNEL PROCESSOR
PCM VOICE TO SWITCH OR PSTN
1701
CHANNEL PROCESSOR
1702
CONTROL PROCESSOR
RECEIVE INTERFERENCE CANCELLATION CONTROL
1703
VOICE OR CONTROL CHANNEL PROCESSOR
PCM VOICE FROM SWITCH OR PSTN
1704
TRANSMIT MATRIX PROCESSOR
DIGITIZED SIGNALS TO HUB R.F. SECTION

FIG. 17

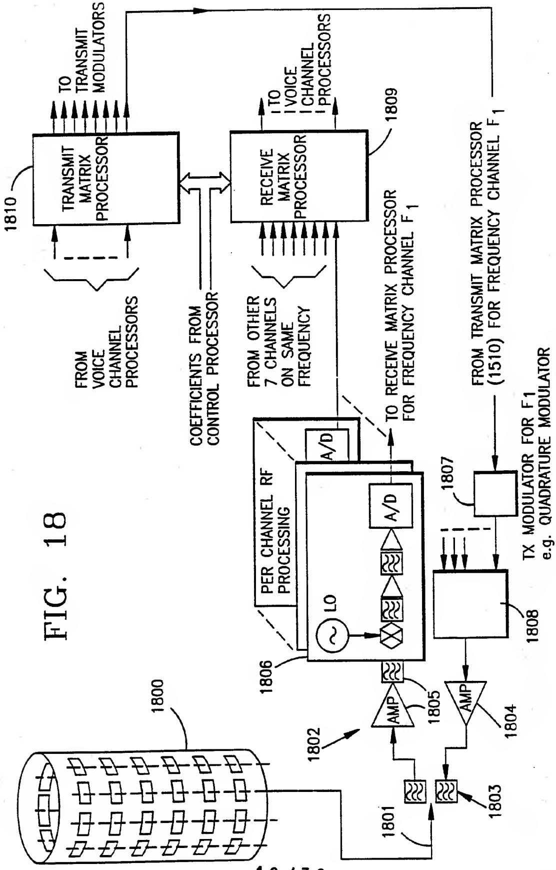

TO TRANSMIT MODULATORS
TRANSMIT MATRIX PROCESSOR
1810
FROM VOICE CHANNEL PROCESSORS
COEFFICIENTS FROM CONTROL PROCESSOR
TO VOICE CHANNEL PROCESSORS
RECEIVE MATRIX PROCESSOR
1809
FROM OTHER 7 CHANNELS ON SAME FREQUENCY
TO RECEIVE MATRIX PROCESSOR FOR FREQUENCY CHANNEL F1
FROM TRANSMIT MATRIX PROCESSOR (1510) FOR FREQUENCY CHANNEL F1
PER CHANNEL RF PROCESSING
A/D
A/D
LO
1806
1807
TX MODULATOR FOR F1 e.g. QUADRATURE MODULATOR
1808
1800
1802
AMP
1805
1805
AMP
1804
1801
1803

FIG. 18

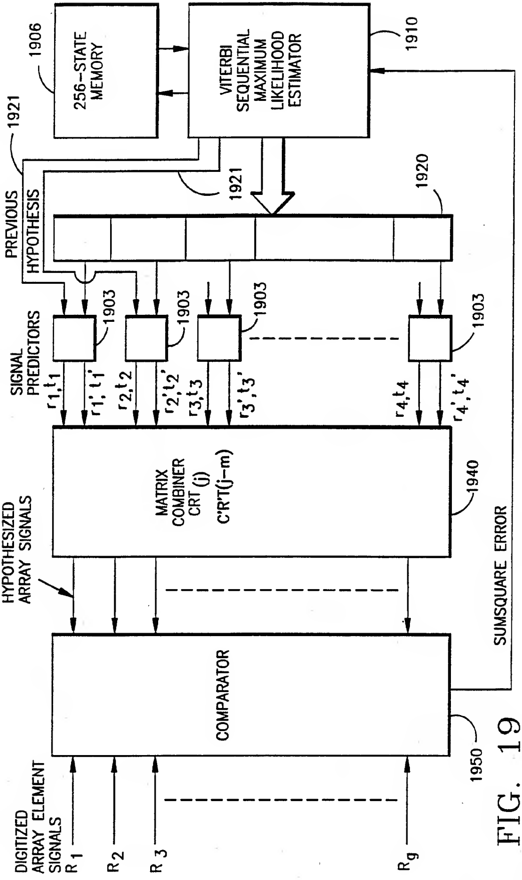
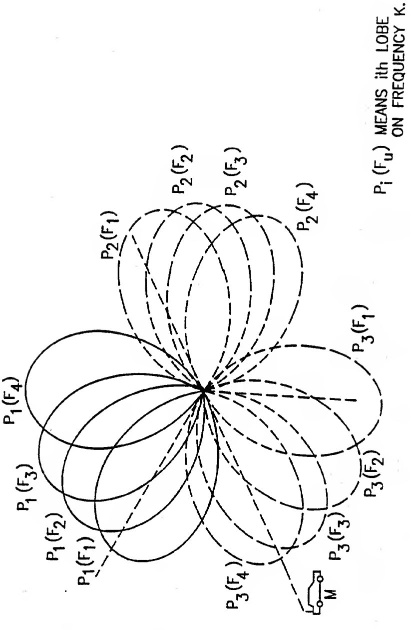

256-STATE MEMORY
1906
VITERBI SEQUENTIAL MAXIMUM LIKELIHOOD ESTIMATOR
1910
1921
PREVIOUS HYPOTHESIS
1921
1920
SIGNAL PREDICTORS
1903
r_1, t_1
r_1', t_1'
r_2, t_2
r_2', t_2'
r_3, t_3
r_3', t_3'
r_4, t_4
r_4', t_4'
MATRIX COMBINER CRT (j) C'R'T(j-m)
1940
HYPOTHESIZED ARRAY SIGNALS
SUMSQUARE ERROR
COMPARATOR
1950
DIGITIZED ARRAY ELEMENT SIGNALS
R_1
R_2
R_3
R_g

FIG. 19



$P_i(F_u)$ MEANS ith LOBE ON FREQUENCY K.
$P_1(F_1)$
$P_1(F_2)$
$P_1(F_3)$
$P_1(F_4)$
$P_2(F_1)$
$P_2(F_2)$
$P_2(F_3)$
$P_2(F_4)$
$P_3(F_1)$
$P_3(F_2)$
$P_3(F_3)$
$P_3(F_4)$
M

FIG. 20

FIG. 21(a)

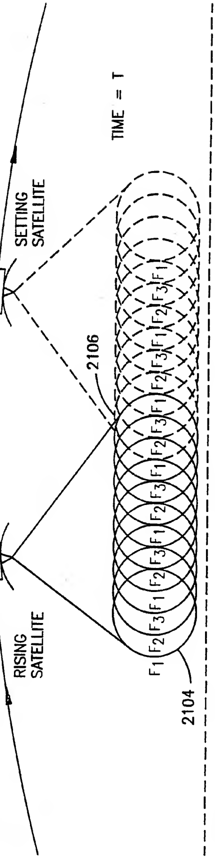


FIG. 21(b)

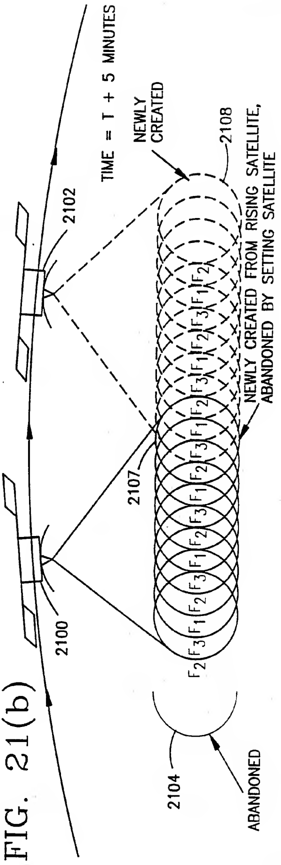
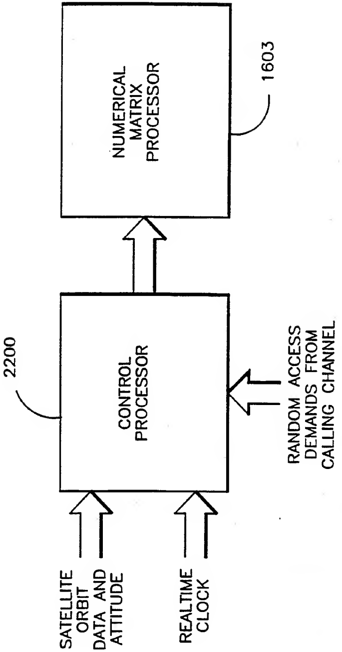


FIG. 22


NUMERICAL MATRIX PROCESSOR
1603
CONTROL PROCESSOR
2200
RANDOM ACCESS DEMANDS FROM CALLING CHANNEL
SATELLITE ORBIT DATA AND ATTITUDE
REALTIME CLOCK

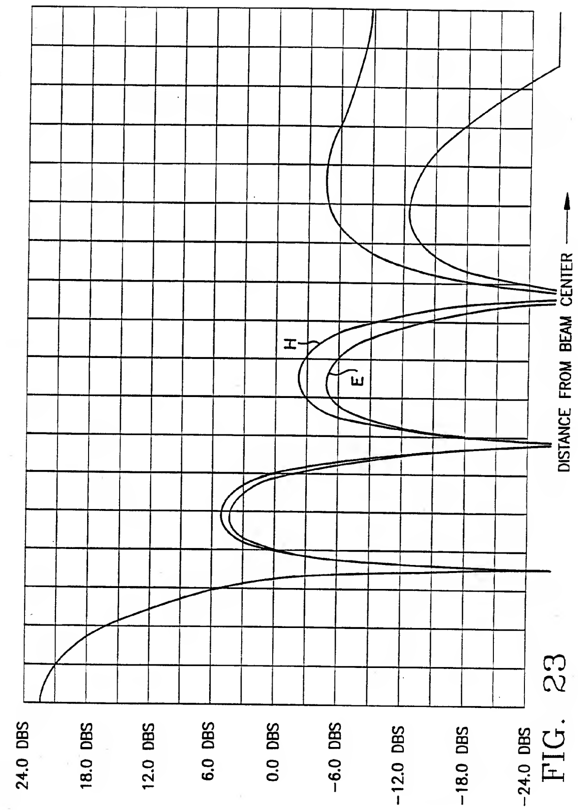

24.0 DBS
18.0 DBS
12.0 DBS
6.0 DBS
0.0 DBS
-6.0 DBS
-12.0 DBS
-18.0 DBS
-24.0 DBS
H
E
DISTANCE FROM BEAM CENTER

FIG. 23

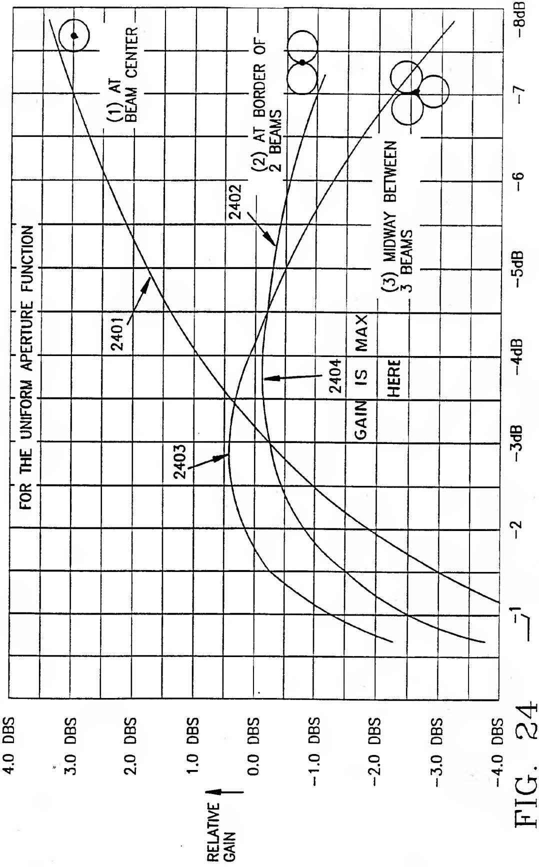

FOR THE UNIFORM APERTURE FUNCTION
(1) AT BEAM CENTER
(2) AT BORDER OF 2 BEAMS
(3) MIDWAY BETWEEN 3 BEAMS
2401
2402
2403
2404
GAIN IS MAX HERE
4.0 DBS
3.0 DBS
2.0 DBS
1.0 DBS
0.0 DBS
-1.0 DBS
-2.0 DBS
-3.0 DBS
-4.0 DBS
RELATIVE GAIN
-1
-2
-3dB
-4dB
-5dB
-6
-7
-8dB

FIG. 24

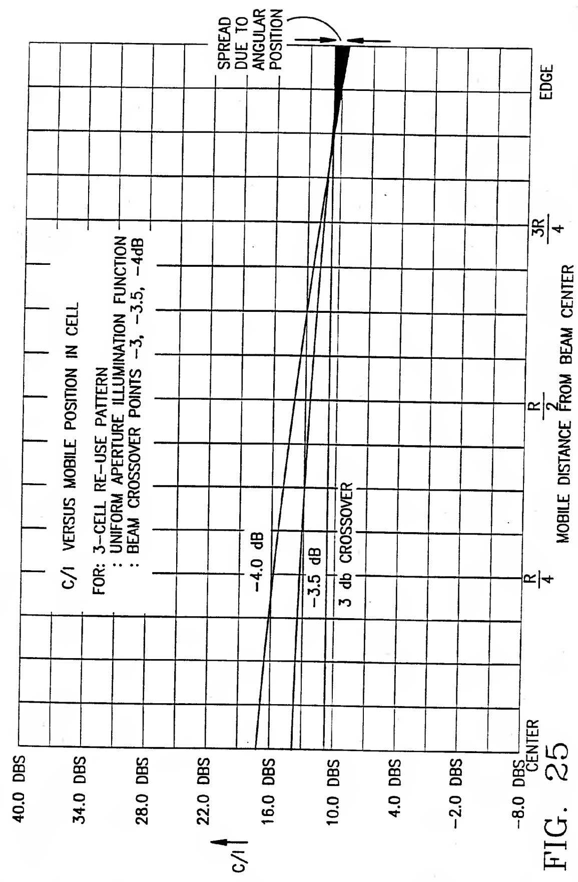

C/I VERSUS MOBILE POSITION IN CELL
FOR: 3-CELL RE-USE PATTERN
: UNIFORM APERTURE ILLUMINATION FUNCTION
: BEAM CROSSOVER POINTS -3, -3.5, -4dB
SPREAD DUE TO ANGULAR POSITION
-4.0 dB
-3.5 dB
3 db CROSSOVER
40.0 DBS
34.0 DBS
28.0 DBS
22.0 DBS
16.0 DBS
10.0 DBS
4.0 DBS
-2.0 DBS
-8.0 DBS
C/I
CENTER
R/4
R/2
3R/4
EDGE
MOBILE DISTANCE FROM BEAM CENTER

FIG. 25

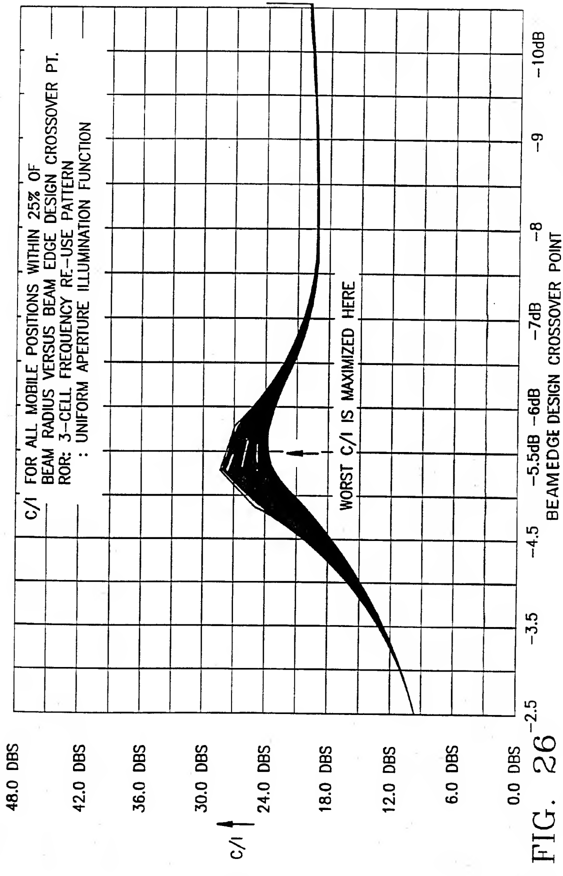

C/I FOR ALL MOBILE POSITIONS WITHIN 25% OF
BEAM RADIUS VERSUS BEAM EDGE DESIGN CROSSOVER PT.
ROR: 3-CELL FREQUENCY RE-USE PATTERN
: UNIFORM APERTURE ILLUMINATION FUNCTION
WORST C/I IS MAXIMIZED HERE
48.0 DBS
42.0 DBS
36.0 DBS
30.0 DBS
24.0 DBS
18.0 DBS
12.0 DBS
6.0 DBS
0.0 DBS
C/I
-2.5
-3.5
-4.5
-5.5dB
-6dB
-7dB
-8
-9
-10dB
BEAMEDGE DESIGN CROSSOVER POINT

FIG. 26

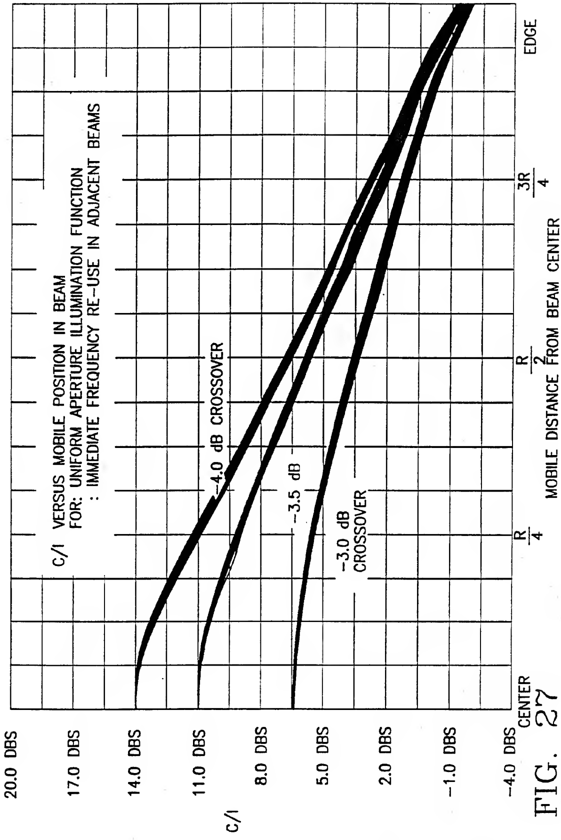

C/I VERSUS MOBILE POSITION IN BEAM
FOR: UNIFORM APERTURE ILLUMINATION FUNCTION
: IMMEDIATE FREQUENCY RE-USE IN ADJACENT BEAMS
-4.0 dB CROSSOVER
-3.5 dB
-3.0 dB CROSSOVER
20.0 DBS
17.0 DBS
14.0 DBS
11.0 DBS
8.0 DBS
5.0 DBS
2.0 DBS
-1.0 DBS
-4.0 DBS
C/I
CENTER
R/4
R/2
3R/4
EDGE
MOBILE DISTANCE FROM BEAM CENTER

FIG. 27

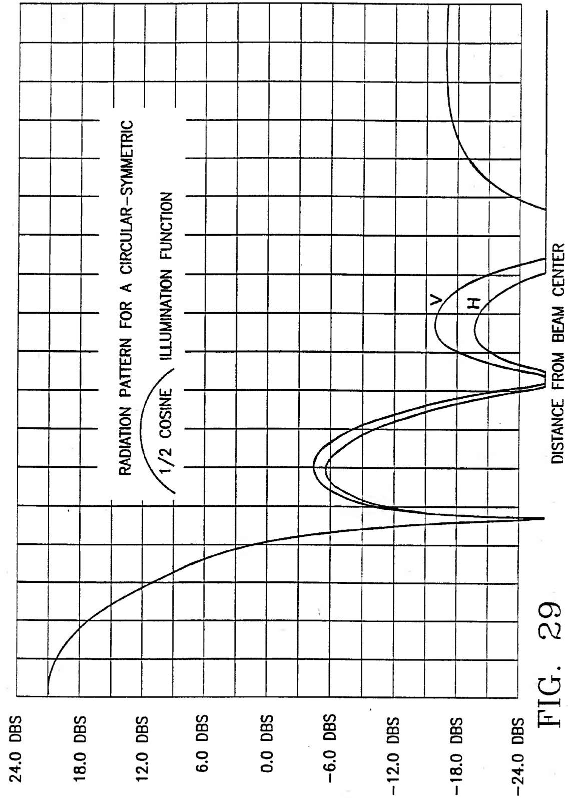

RADIATION PATTERN FOR A CIRCULAR-SYMMETRIC
1/2 COSINE ILLUMINATION FUNCTION
V
H
DISTANCE FROM BEAM CENTER
24.0 DBS
18.0 DBS
12.0 DBS
6.0 DBS
0.0 DBS
-6.0 DBS
-12.0 DBS
-18.0 DBS
-24.0 DBS

FIG. 29

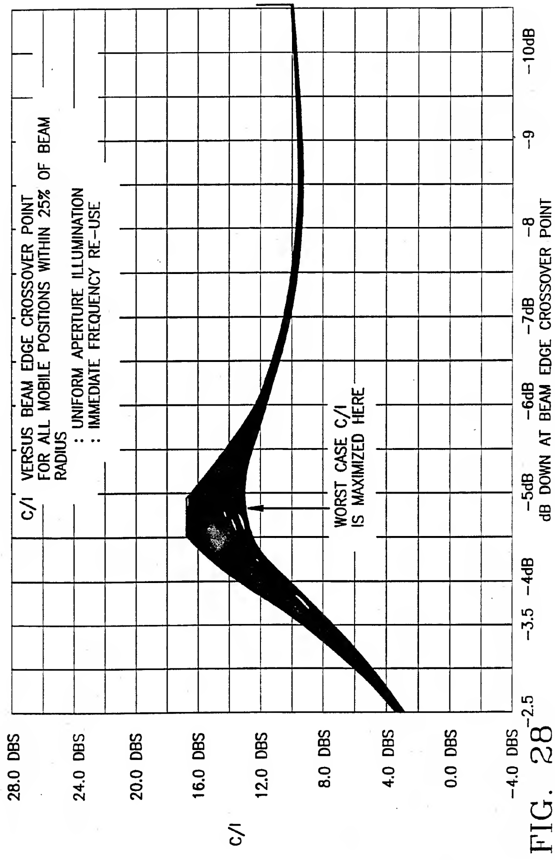

C/I VERSUS BEAM EDGE CROSSOVER POINT
FOR ALL MOBILE POSITIONS WITHIN 25% OF BEAM RADIUS
: UNIFORM APERTURE ILLUMINATION
: IMMEDIATE FREQUENCY RE-USE
WORST CASE C/I
IS MAXIMIZED HERE
28.0 DBS
24.0 DBS
20.0 DBS
16.0 DBS
12.0 DBS
8.0 DBS
4.0 DBS
0.0 DBS
-4.0 DBS
C/I
-2.5
-3.5
-4dB
-5dB
-6dB
-7dB
-8
-9
-10dB
dB DOWN AT BEAM EDGE CROSSOVER POINT

FIG. 28

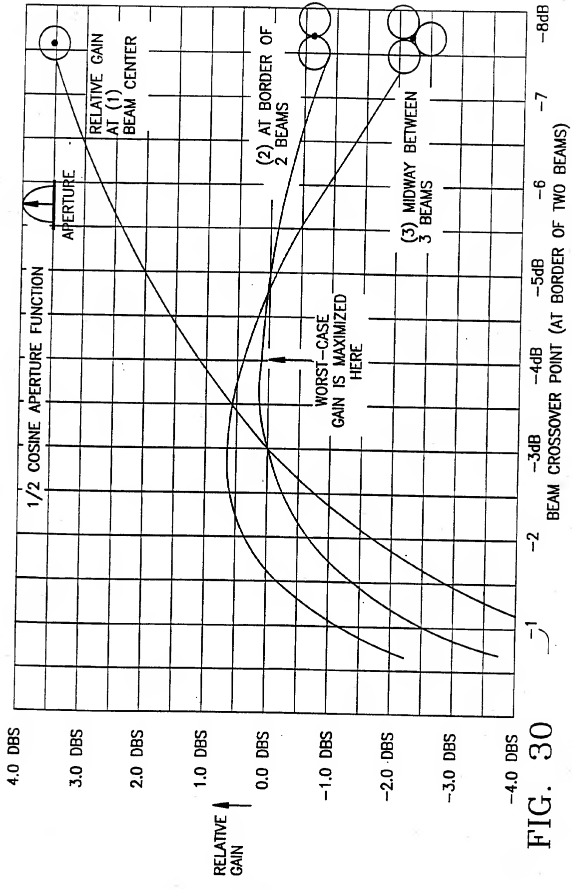

1/2 COSINE APERTURE FUNCTION
APERTURE
RELATIVE GAIN AT (1) BEAM CENTER
(2) AT BORDER OF 2 BEAMS
(3) MIDWAY BETWEEN 3 BEAMS
WORST-CASE GAIN IS MAXIMIZED HERE
4.0 DBS
3.0 DBS
2.0 DBS
1.0 DBS
0.0 DBS
-1.0 DBS
-2.0 DBS
-3.0 DBS
-4.0 DBS
RELATIVE GAIN
-1
-2
-3dB
-4dB
-5dB
-6
-7
-8dB
BEAM CROSSOVER POINT (AT BORDER OF TWO BEAMS)

FIG. 30

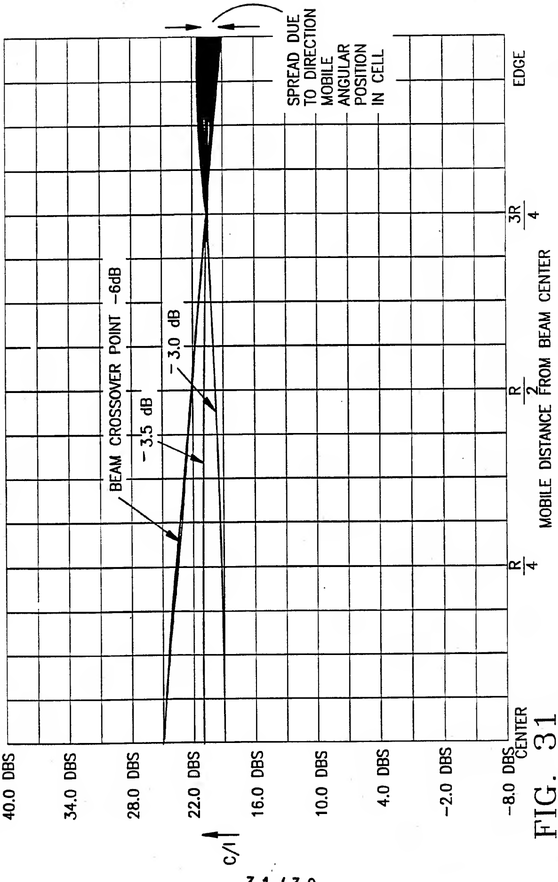

SPREAD DUE TO DIRECTION MOBILE ANGULAR POSITION IN CELL
BEAM CROSSOVER POINT −6dB
− 3.5 dB
− 3.0 dB
40.0 DBS
34.0 DBS
28.0 DBS
22.0 DBS
16.0 DBS
10.0 DBS
4.0 DBS
−2.0 DBS
−8.0 DBS
C/I
CENTER
R/4
R/2
3R/4
EDGE
MOBILE DISTANCE FROM BEAM CENTER

FIG. 31

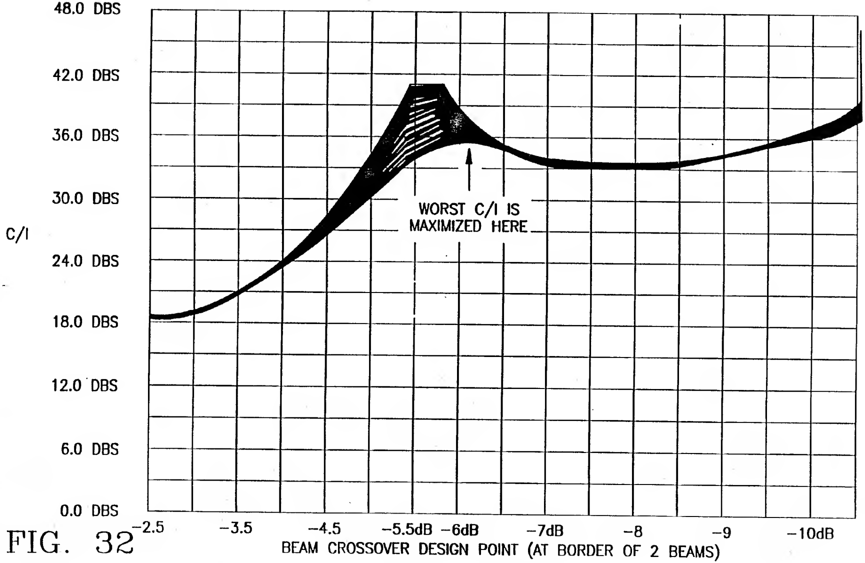
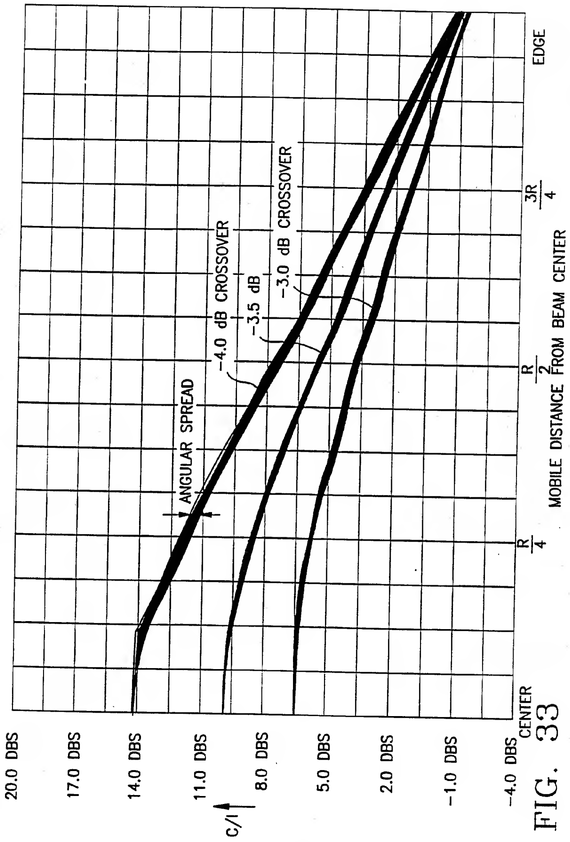

48.0 DBS
42.0 DBS
36.0 DBS
30.0 DBS
C/I
24.0 DBS
18.0 DBS
12.0 DBS
6.0 DBS
0.0 DBS
WORST C/I IS MAXIMIZED HERE
-2.5
-3.5
-4.5
-5.5dB -6dB
-7dB
-8
-9
-10dB
BEAM CROSSOVER DESIGN POINT (AT BORDER OF 2 BEAMS)

FIG. 32



20.0 DBS
17.0 DBS
14.0 DBS
11.0 DBS
8.0 DBS
5.0 DBS
2.0 DBS
-1.0 DBS
-4.0 DBS
C/I
CENTER
R/4
R/2
3R/4
EDGE
MOBILE DISTANCE FROM BEAM CENTER
ANGULAR SPREAD
-4.0 dB CROSSOVER
-3.5 dB
-3.0 dB CROSSOVER

FIG. 33

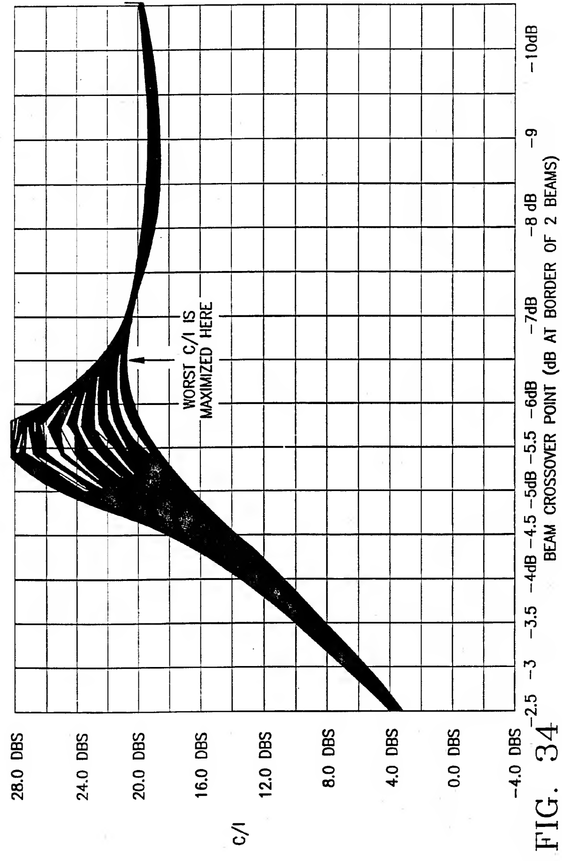
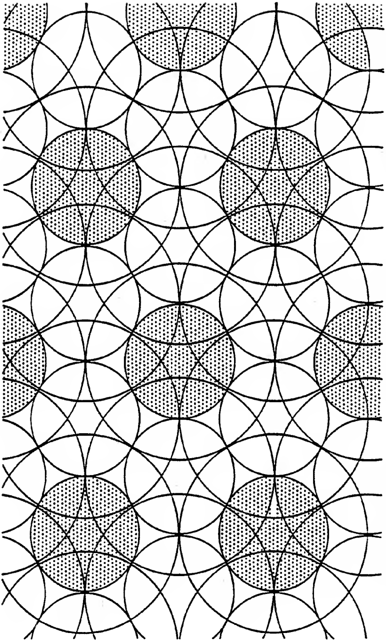
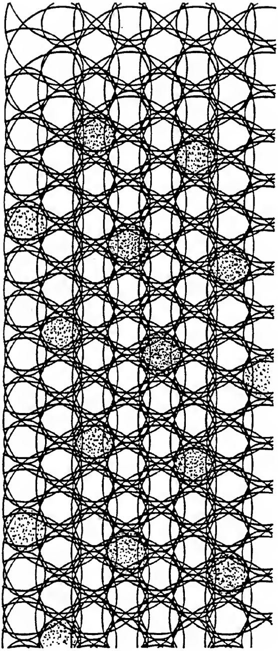

28.0 DBS
24.0 DBS
20.0 DBS
16.0 DBS
12.0 DBS
8.0 DBS
4.0 DBS
0.0 DBS
−4.0 DBS
C/I
−2.5 −3 −3.5 −4dB −4.5 −5dB −5.5 −6dB −7dB −8 dB −9 −10dB
BEAM CROSSOVER POINT (dB AT BORDER OF 2 BEAMS)
WORST C/I IS MAXIMIZED HERE

FIG. 34



THE CHARACTERS OF EACH GIVEN COLOR ARE TOUCHING (1 CELL PLAN)
BUT THE SMALLER CIRCLES FORM A 3-CELL PATTERN

FIG. 35



7 COLORS USED SYSTEMATICALLY SUCH THAT LARGER CIRCLES OF ANY COLOR TOUCH WHILE THE SMALLER CIRCLES FORM A 7-CELL RE-USE PATTERN WITH THE OTHER COLORS

FIG. 36

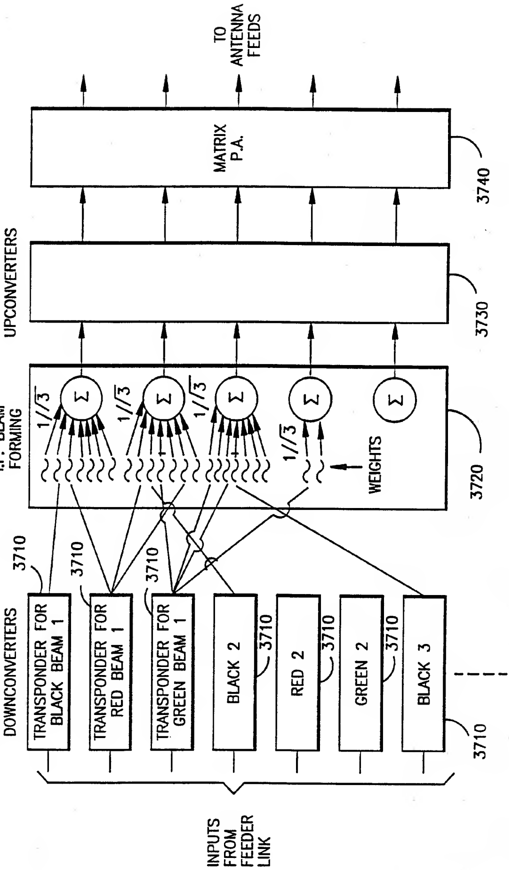
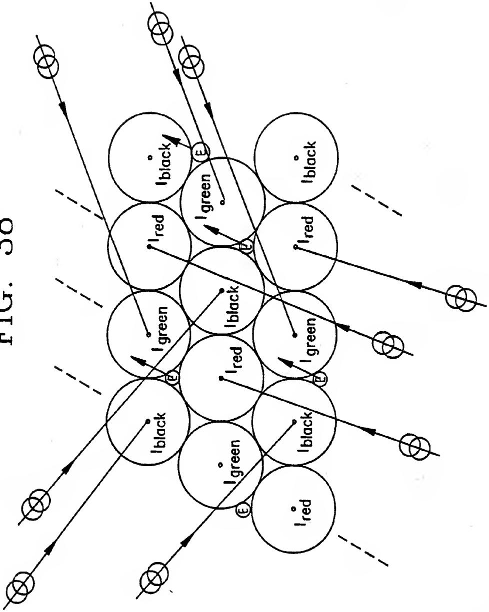

TO ANTENNA FEEDS
MATRIX P.A.
3740
UPCONVERTERS
3730
I.F. BEAM FORMING
1/√3
WEIGHTS
3720
DOWNCONVERTERS
TRANSPONDER FOR BLACK BEAM 1
TRANSPONDER FOR RED BEAM 1
TRANSPONDER FOR GREEN BEAM 1
BLACK 2
RED 2
GREEN 2
BLACK 3
3710
INPUTS FROM FEEDER LINK

FIG. 37

FIG. 38



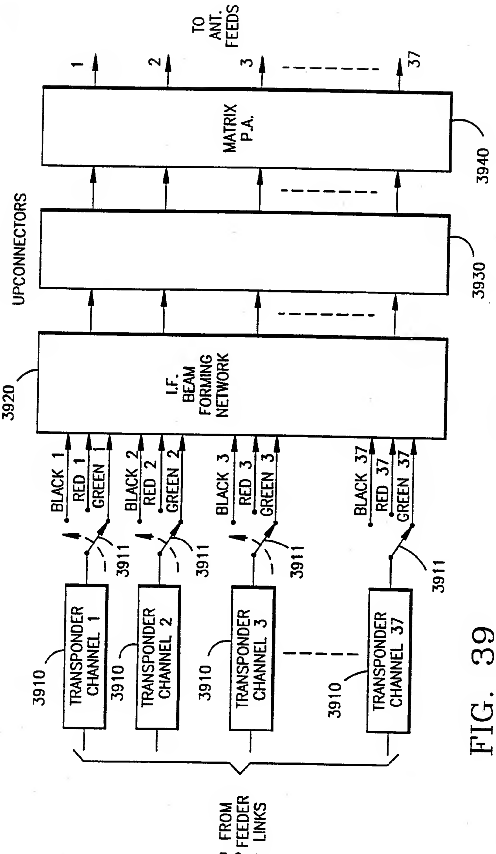

TO ANT. FEEDS
1
2
3
37
MATRIX P.A.
3940
UPCONNECTORS
3930
3920
I.F. BEAM FORMING NETWORK
BLACK 1
RED 1
GREEN 1
BLACK 2
RED 2
GREEN 2
BLACK 3
RED 3
GREEN 3
BLACK 37
RED 37
GREEN 37
3911
3911
3911
3911
3910
TRANSPONDER CHANNEL 1
3910
TRANSPONDER CHANNEL 2
3910
TRANSPONDER CHANNEL 3
3910
TRANSPONDER CHANNEL 37
FROM FEEDER LINKS

FIG. 39

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US95/00224

A. CLASSIFICATION OF SUBJECT MATTER

IPC(6) :H04J 3/16; H04B 1/04, 1/16, 7/26, 7/185; H04Q 3/04
US CL : 370/95.3; 455/12.1, 13.1, 13.4, 38.1, 54.1
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

U.S. : 370/32, 35, 75, 95.1, 95.3, 104.1, 121, 123; 455/12.1, 13.1, 13.2, 13.4, 17, 22, 38.1, 38.3, 49.1, 53.1, 54.1

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A, E	US, A, 5,398,247 (DELPRAT ET AL) 14 MARCH 1995	1-13, 17-22, 26-30, 44-68, 77-79
A	US, A, 4,907,003 (MARSHALL ET AL) 06 MARCH 1990	14-16, 76
A	US, A, 5,247,702 (SU ET AL) 21 SEPTEMBER 1993	23-25, 31-35, 38
A	US, A, 4,876,737 (WOODWORTH ET AL) 24 OCTOBER 1989	36-37, 39-43
A	US, A, 4,752,967 (BUSTAMANTE ET AL) 21 JUNE 1988	69-75

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

* Special categories of cited documents:

"A" document defining the general state of the art which is not considered to be part of particular relevance

"E" earlier document published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
24 APRIL 1995	26 MAY 1995

Name and mailing address of the ISA/US	Authorized officer
Commissioner of Patents and Trademarks Box PCT Washington, D.C. 20231	RICKY Q. NGO
Facsimile No. (703) 305-3230	Telephone No. (703) 305-4798

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US95/00224

Box I Observations where certain claims were found unsearchable (Continuation of item 1 of first sheet)

This international report has not been established in respect of certain claims under Article 17(2)(a) for the following reasons:

1. ☐ Claims Nos.:
   because they relate to subject matter not required to be searched by this Authority, namely:

2. ☐ Claims Nos.:
   because they relate to parts of the international application that do not comply with the prescribed requirements to such an extent that no meaningful international search can be carried out, specifically:

3. ☐ Claims Nos.:
   because they are dependent claims and are not drafted in accordance with the second and third sentences of Rule 6.4(a).

Box II Observations where unity of invention is lacking (Continuation of item 2 of first sheet)

This International Searching Authority found multiple inventions in this international application, as follows:

Please See Extra Sheet.

1. ☒ As all required additional search fees were timely paid by the applicant, this international search report covers all searchable claims.

2. ☐ As all searchable claims could be searched without effort justifying an additional fee, this Authority did not invite payment of any additional fee.

3. ☐ As only some of the required additional search fees were timely paid by the applicant, this international search report covers only those claims for which fees were paid, specifically claims Nos.:

4. ☐ No required additional search fees were timely paid by the applicant. Consequently, this international search report is restricted to the invention first mentioned in the claims; it is covered by claims Nos.:

Remark on Protest ☐ The additional search fees were accompanied by the applicant's protest.

☒ No protest accompanied the payment of additional search fees.

BOX II. OBSERVATIONS WHERE UNITY OF INVENTION WAS LACKING
This ISA found multiple inventions as follows:

This application contains the following inventions or groups of inventions which are not so linked as to form a single inventive concept under PCT Rule 13.1. In order for all inventions to be examined, the appropriate additional examination fees must be paid.

Group I. Claims 1-13, 17-22, 26-30, 44-68 and 77-79, drawn to Time Division Multiple Access, classified in Class 370, subclass 95.3.

Group II. Claims 14-16 and 76, drawn to space satellite with antenna feed network or multiple antenna switching, classified in Class 455, subclass 13.3.

Group III. Claims 23-25, 31-35 and 38, drawn to system with receiver selection with coded sequence, classified in class 455, subclass 38.1.

Group IV. Claims 36-37, 39-43, drawn to satellite transponder transmitting received signals with a new carrier frequency, classified in Class 455, subclass 12.1.

Group V. Claims 69-74, drawn to space satellite with power control, classified in Class 455, subclass 13.4.

Group VI. Claim 75, drawn to wireless communications between a base station and a plurality of mobile stations, classified in Class 455, subclass 54.1.The inventions listed above as Groups do not relate to a single inventive concept under PCTRule13.1because,under PCT Rule 13.2, they lack the same or corresponding special technical features for the following reasons: The inventions listed as Groups I-VI do not relate to a single inventive concept under PCT Rule 13.1 because, under PCT Rule 13.2, they lack the same or corresponding special technical features for the following reasons: Groups I-VI each have different modes of operation, and they have different functions.